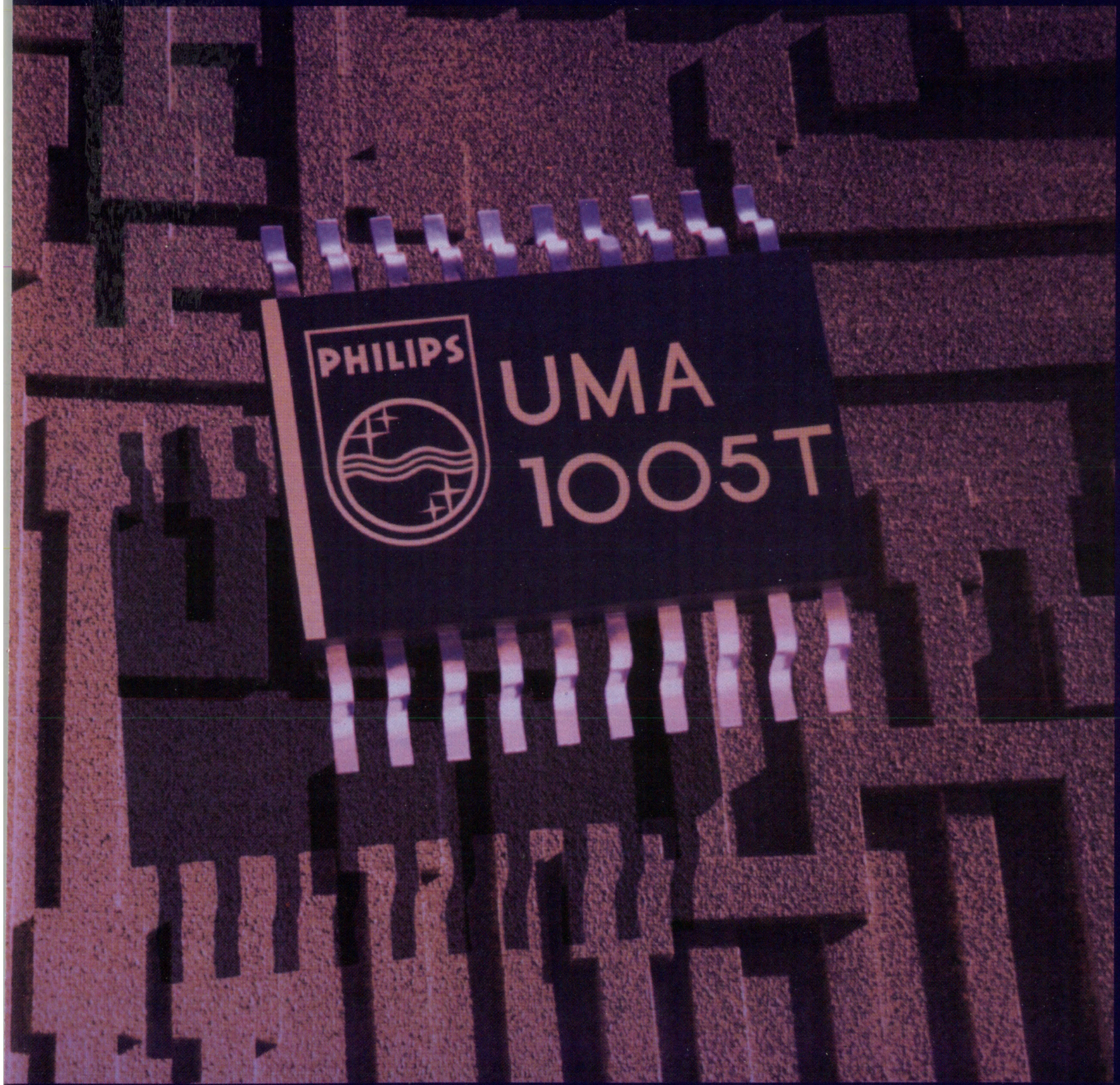
# RF design

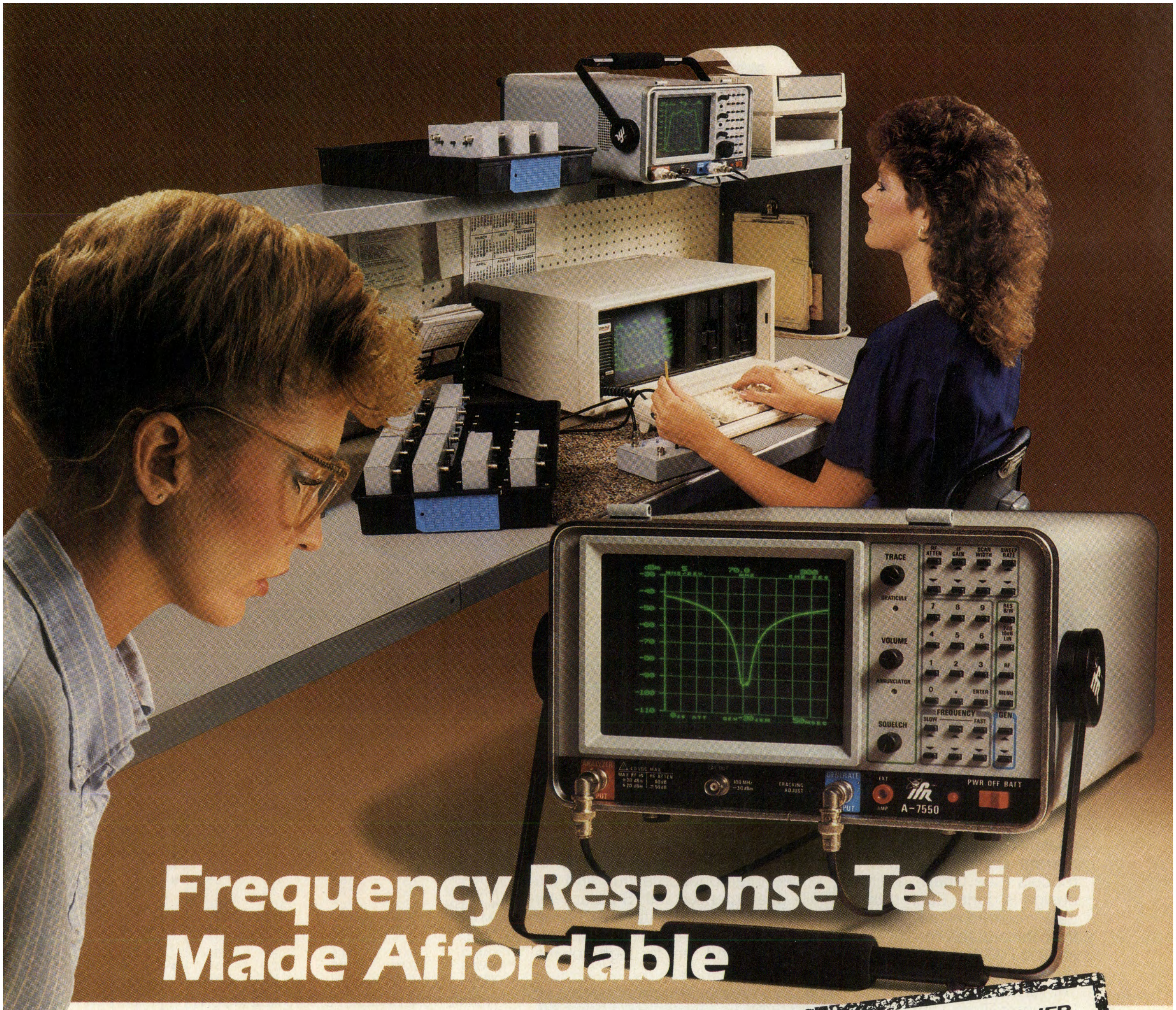
engineering principles and practices

February 1993




*Cover Story*
**Fractional-N Synthesizer IC Simplifies Design**

*Plus —*
**RF Expo West Technical Program Abstracts**

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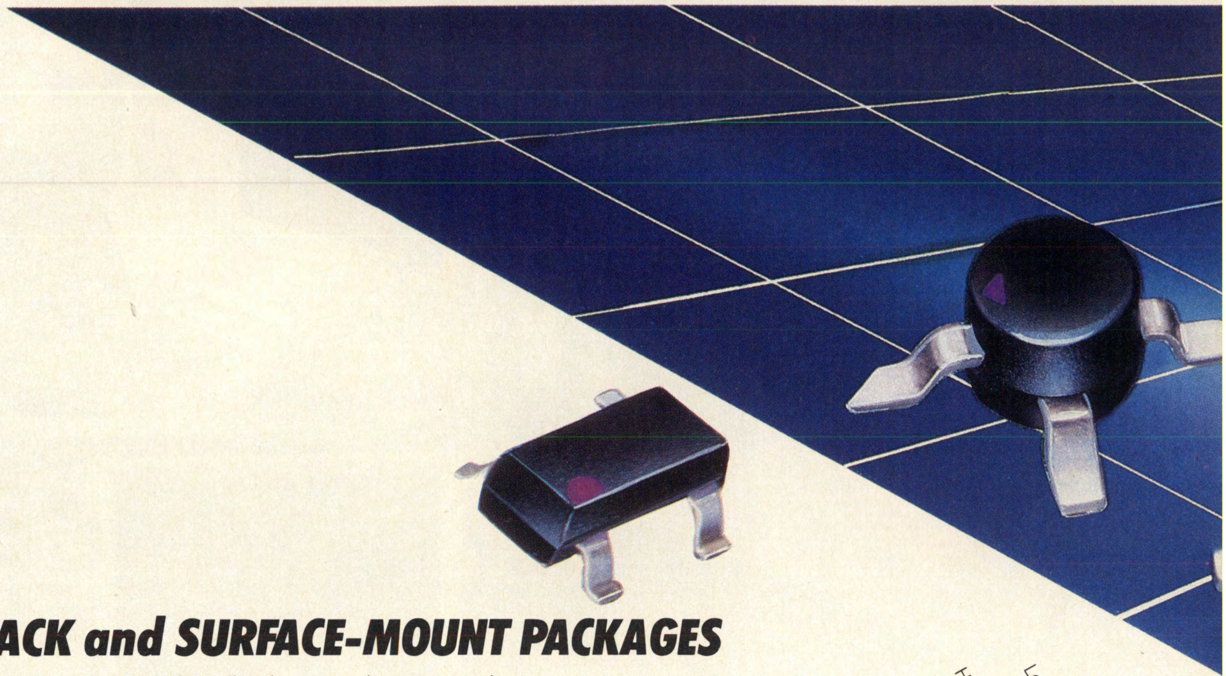
INFO/CARD 2

# dc to 2000 MHz



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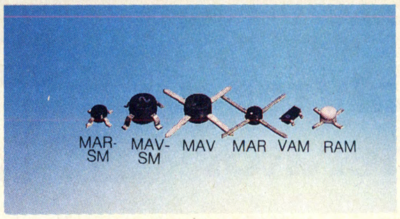
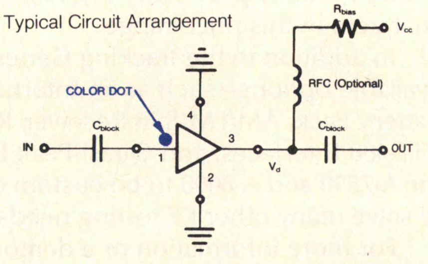



Models above shown actual size

Typical Circuit Arrangement




SPECIFICATIONS

| Model No. | Freq.,MHz dc to: | Gain,dB 100 MHz | Output Pwr +dBm | NF dB | High Gain | Power | Low Noise | Price, $ ea. (25 Qty) |
|---|---|---|---|---|---|---|---|---|
| | | | | | (features) | | | |
| **SURFACE MOUNT** | | | | | | | | |
| MAR-1SM | 1000 | 18.5 | 1.5 | 5.5 | ● | | | 1.04 |
| MAR-2SM | 2000 | 12.5 | 4.5 | 6.5 | | | | 1.40 |
| MAR-3SM | 2000 | 12.5 | 10.0 | 6.0 | | | | 1.50 |
| MAR-4SM | 1000 | 8.3 | 12.5 | 7.0 | | ● | | 1.60 |
| MAR-6SM | 2000 | 20.0 | 2.0 | 3.0 | ● | | ● | 1.34 |
| MAR-7SM | 2000 | 13.5 | 5.5 | 5.0 | | | | 1.80 |
| MAR-8SM | 1000 | 32.5 | 12.5 | 3.3 | ● | ● | ● | 1.75 |
| MAV-1SM | 1000 | 18.5 | 1.5 | 5.5 | ● | | | 1.15 |
| MAV-2SM | 1500 | 12.5 | 4.5 | 6.5 | | | | 1.45 |
| MAV-3SM | 1500 | 12.5 | 10.0 | 6.0 | | | | 1.55 |
| MAV-4SM | 1000 | 8.3 | 11.5 | 7.0 | | | | 1.65 |
| MAV-11SM | 1000 | 12.7 | 17.5 | 3.6 | | ● | ● | 2.15 |
| VAM-3 | 2000 | 11.5 | 9.0 | 6.0 | | | | 1.45 |
| VAM-6 | 2000 | 19.5 | 2.0 | 3.0 | ● | | ● | 1.29 |
| VAM-7 | 2000 | 13.0 | 5.5 | 5.0 | | | | 1.75 |
| **HERMETICALLY-SEALED SURFACE MOUNT** | | | | | | | | |
| RAM-1 | 1000 | 19.0 | 1.5 | 5.5 | ● | | | 4.95 |
| RAM-2 | 2000 | 12.5 | 4.5 | 6.5 | | | | 4.95 |
| RAM-3 | 2000 | 12.5 | 10.0 | 6.0 | | | | 4.95 |
| RAM-4 | 1000 | 8.5 | 12.5 | 6.5 | | ● | | 4.95 |
| RAM-6 | 2000 | 20.0 | 2.0 | 2.8 | ● | | ● | 4.95 |
| RAM-7 | 2000 | 13.5 | 5.5 | 4.5 | | | | 4.95 |
| RAM-8 | 1000 | 32.5 | 12.5 | 3.0 | ● | ● | ● | 4.95 |
| **FLATPACK** | | | | | | | | |
| MAR-1 | 1000 | 18.5 | 1.5 | 5.5 | ● | | | 0.99 |
| MAR-2 | 2000 | 12.5 | 4.5 | 6.5 | | | | 1.35 |
| MAR-3 | 2000 | 12.5 | 10.0 | 6.0 | | | | 1.45 |
| MAR-4 | 1000 | 8.3 | 12.5 | 6.5 | | ● | | 1.55 |
| MAR-6 | 2000 | 20.0 | 2.0 | 3.0 | ● | | ● | 1.29 |
| MAR-7 | 2000 | 13.5 | 5.5 | 5.0 | | | | 1.75 |
| MAR-8 | 1000 | 32.5 | 12.5 | 3.3 | ● | ● | ● | 1.70 |
| MAV-1 | 1000 | 18.5 | 1.5 | 5.5 | ● | | | 1.10 |
| MAV-2 | 1500 | 12.5 | 4.5 | 6.5 | | | | 1.40 |
| MAV-3 | 1500 | 12.5 | 10.0 | 6.0 | | | | 1.50 |
| MAV-4 | 1000 | 8.3 | 11.5 | 7.0 | | | | 1.60 |
| MAV-11 | 1000 | 12.7 | 17.5 | 3.6 | | ● | ● | 2.10 |

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***DAK-2SM:*** 5 of each MAR-SM model (35 pc), only $61.95

***DAK-3:*** 3 of each MAR-, -SM, MAV-11, -11SM (48 pc), $59.95

designer's capacitor kit

***KCAP-1:*** 50 of 17 values, 10pf to 0.1uf (850 pc), $99.95

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F 147 REV. A

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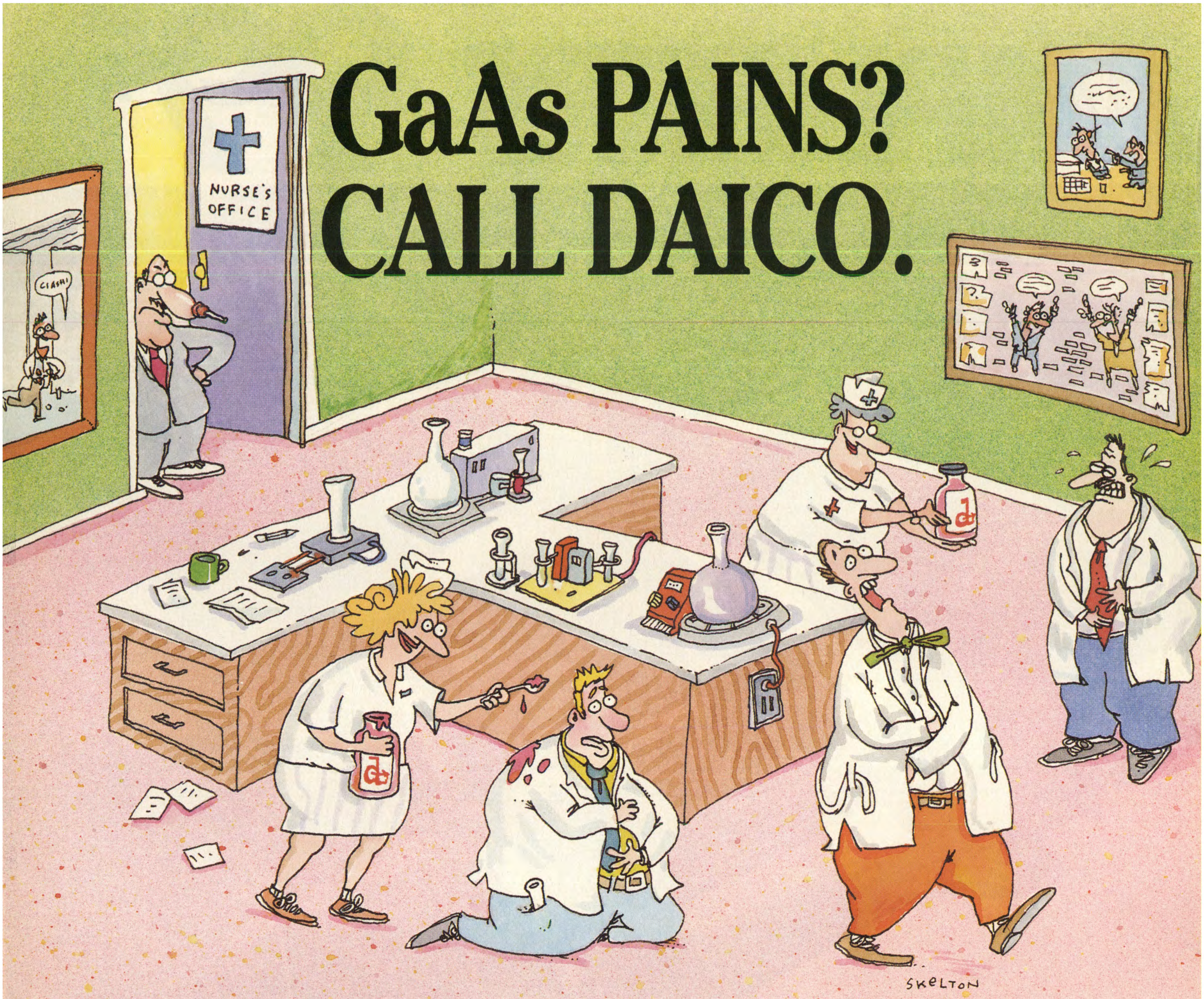


### Stock GaAs Bi-Phase Modulators

| Freq | IL dB | Switch Speed nSEC | Phase Accuracy Degree | Con-trol | Package | Part No. |
|---|---|---|---|---|---|---|
| 10-500 MHz | 1.0 | 50 | +/-1 | TTL | TO8 | DBP0938 |
| 10-500 MHz | 1.0 | 50 | +/-1 | TTL | SMP | DBP0738 |

### Stock Packaged GaAs Switches

| Con-fig | Freq MHz | IL dB | Iso dB | Switch Speed nSEC | Con-trol | Package | Part No. |
|---|---|---|---|---|---|---|---|
| SPST | 5-1500 | 1.0 | 48 | 25 | TTL | TO-8 | DSO699 |
| SPST | 10-2000 | 1.8 | 67 | 35 | TTL | 14 Pin SMP | DSO790 |
| SP2T | DC-2000 | 0.5 | 35 | 3 | – | 8 Pin SOIC | DSO702R |
| SP2T | DC-2000 | 0.6 | 30 | 3 | – | 8 Pin SOIC | DSO702T |
| SP2T | DC-2000 | 0.4 | 68 | 3 | – | TO-5 | DSO850 |
| SP2T | DC-2000 | 0.7 | 50 | 200 | TTL | TO-5 | DSO813 |
| SP2T | 5-2000 | 1.15 | 55 | 35 | TTL | 14 Pin DIP | DSO602 |
| SP2T | 5-4000 | 1.0 | 79 | 35 | TTL | SMA | CDSO882 |
| SP4T | DC-2000 | 1.7 | 70 | 75 | TTL | 14 Pin DIP | DSO874 |

### Stock GaAs MMIC Amplifiers

| Freq Range/GHz | Typ Gain/dB | Typ Noise Figure | Typ Power /dBm | Package | Comments | Part No. |
|---|---|---|---|---|---|---|
| 0.05-3.5 | 10 | 6.0 | 22 | Chip | | P35-4100-0 |
| 0.5-3.5 | 9 | 4.5 | 22 | Chip | Self-Biased | P35-4101-0 |
| 0.05-3.0 | 18 | 6.0 | 13 | Chip | Low VSWRs | P35-4104-0 |
| 0.8-1.8 | 21 | 3.5 | 8 | Chip | | P35-4105-0 |
| 1-6 | 7.5 | 4.6 | 20 | Chip | | P35-4110-0 |
| 6-18 | 5.5 | 5.5 | 15 | Chip | Pos.Gain Slope | P35-4140-0 |
| 2-18 | 6.0 | 7.5 | 15 | Chip | AGC | P35-4150-0 |
| 3-6 | 2.0 | 2.8 | 14 | Chip | Low VSWRs | P35-4160-0 |

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# RF design

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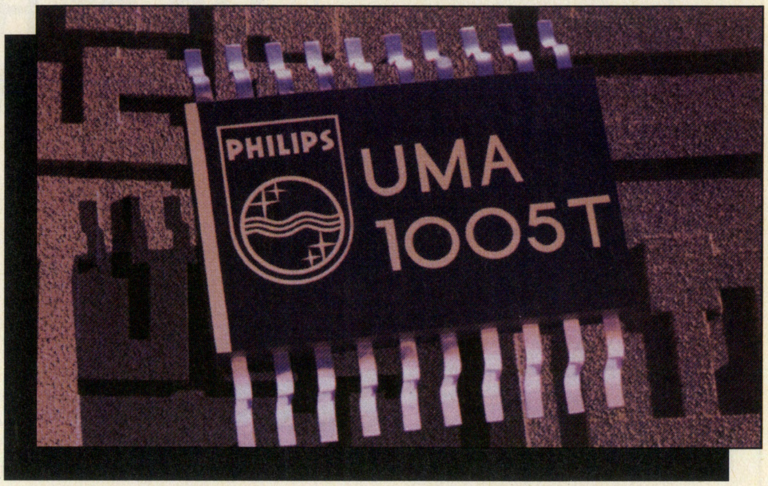


## featured technology



## cover story



## tutorial



## design awards



## departments



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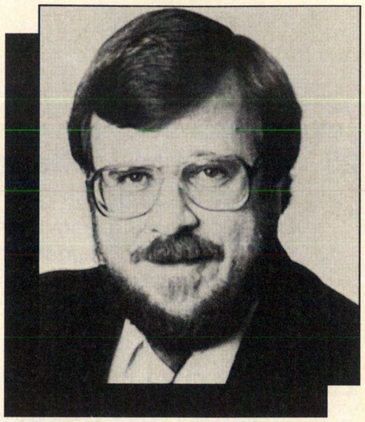
*RF editorial*

# Come See Us at RF Expo West



*By Gary A. Breed*
*Editor*

Do you need a few good reasons to come to RF Expo West next month (March 17-19)? Well, let's start with our special courses:

*Introduction to RF Circuit Design (two days).* Part I covers RF system and component fundamentals, including definitions of key RF concepts: gain, bandwidth, linear and non-linear, amplifier classes, etc. Transmission line concepts and impedance matching round out the first of two one-day sessions. Part II is the active circuits class, covering a lot of ground from basic component models to computer-aided design, from s-parameter concepts to amplifiers and oscillators. Large-signal circuits and the family of couplers and hybrids wrap up day two. This pair of courses is taught by new RF Expo instructors, Drs. Dave Hertling and Bob Feeney from Georgia Tech. They should be familiar to *RF Design* readers, having written articles and lent their expertise in other ways since the magazine started in 1978. The participation of Georgia Tech also means that we will be able to offer CEU credits for completion of the classes.

*Filters and Matching Networks.* Randy Rhea of Eagleware repeats his updated class on filters and matching networks again in San Jose. This class is now offered without another class competing for attention, so more RF engineers can improve their knowledge of this essential subject. Randy's *Oscillator Design Principles* course is also on the schedule, covering the analysis and design of oscillators circuits in a unified, systematic manner. Both of these classes have received well over 90 percent approval ratings from past students.

If that's not enough, how about these topics that will be covered in the technical papers:

GSM transceivers, Cellular repeaters, Sigma-delta technology, PCN antennas, Ferrite rod antennas, Coaxial cable, Circuit analysis methods, Power amplifier analysis, High efficiency amplifiers, Infrared transmission, Wireless medical network, Miniaturization techniques, Class B/D dual mode amplifier, Low cost class E amplifier, Amplifier impedance measurements, Modulating SAW oscillators, Computer simulation comparison, Circuit radiation modeling, Simulating discontinuities, 1.5-2.0 GHz power transistors, 250 watt amplifier for 100-400 MHz, 50 watt L-band power amplifier, Satellite applications, Simulating transmission impairments, ASICs for RF applications, Miniature circuit examples, PCB test probes, Automated RF testing, DECT modulator/demodulator, Vector modulator, Wideband PLL design, Frequency-independent phase tracking, Knowledge-based systems. Abstracts of these papers are in this issue on page 71.

Finally, we have 160+ companies in the exhibition hall with the latest, greatest and most exciting RF products. You can get up to date on current technology face-to-face with the marketing and applications engineers from these suppliers of instruments, components, software and services. Many exhibitors will have demonstration units operating, so you can get a good hands-on evaluation of their products.

RF Expo West is the first (and is still the best) forum for RF engineers to learn and share ideas about their craft, and the best place to see hundreds of RF products "up close and personal" in the exhibit hall. I'll look forward to seeing you in San Jose!

# RF design

a Cardiff publication Established 1978

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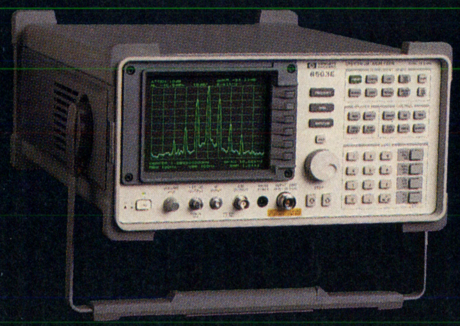

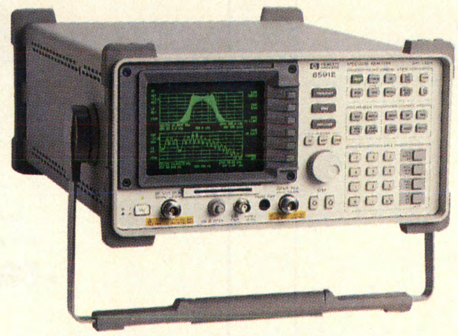

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# RF letters

Letters should be addressed to: Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Published letters may be edited for length or clarity.

### Inductive Coupling *is* Practical

Editor:

I would like to suggest one correction to Andrzej Przedpelski's otherwise excellent tutorial article on double-tuned circuits. (*RF Design*, Nov. 1992, page 52). In the first paragraph under the heading *Capacitive Coupling* the author states:

"In some situations it is desirable to have the possibility of changing the coupling between the two circuits. The inductively coupled configurations (with the exception of Figure 1(c)) are not suitable for this application."

I believe that I can show that not only is it possible to utilize the configuration of Figure 1(a) for this purpose, but that it may be the BEST way, and that it has been done commercially.

The *Radiotron Designer's Handbook*, 3rd. ed. (1941) states on page 158 under the heading, *Variable Selectivity*:

"Variable coupling by means of pure mutual inductance is the only system in which mistuning of the transformer does not occur. As the coupling is increased above the critical value, the "trough" of the resonance curve remains at the intermediate frequency. ( See C.B. Aiken, "Two-mesh Tuned Coupled Circuit Filters," *Proc. I.R.E.*, pg. 230, Feb. 1937)"

If my memory serves me correctly, this system was used commercially in the Hammarlund Super Pro, one of the finest communications receivers available in the 1940's. In this receiver, the primary and secondary windings of the IF transformers were mounted in-line but in such a way that the spacing between them could be changed by a cam arrangement connected to a front panel control. The cam was designed in such a way that the coefficient of coupling would be slightly greater than critical at one limit and, at the other, a degree of undercoupling which would result in the best compromise between the loss of gain and increase in selectivity.

It may be true that modern techniques have eliminated the necessity of this system, but it is not true that the circuit shown in Figure 1(a) is impracticable.

Robert C. Skar
Glen Ellyn, IL

### Turn Down the Power Before Turning Up the Price

Editor:

While I support your idea ("An EMC Wish List", *RFD*, August 1992) that consumer equipment should be more resistant to EMI than most of it presently is, my jaw dropped when I read that the FCC had measured a 9 V/meter field induced in a hapless neighbor's home by a ham running a 1 kW HF transmitter. This is a *huge*, absurd amount of RF for a piece of consumer equipment to reject!

EMI suppression is not free. If a $300 (retail) VCR can have no more than about $40 in actual component cost, the cost of adding sufficient EMI suppression to reject a 9 V/m field can significantly affect the selling price of such equipment. I would ask you why tens of millions of consumers should be so taxed to permit a few amateurs to indulge their hobby. In the 1930's, it did not seem unreasonable to let people fire up 1 kW rigs in residential neighborhoods. In the 90's, it seems absurd to permit this. Reducing power to 10 W would reduce the electric field 20 dB (to 0.9 V/m). While this is still a lot of RF, the power level now becomes comparable to other services, like cellular phones and the like. It is time for the FCC to act to reduce the permissible RF fields that amateurs can blast into their neighbors' homes.

Robert Orban
Belmont, CA

# RF calendar

## February

**9-11**

**The Second Pan-European ESD Symposium and Exhibition**
Strasbourg, France
Information: Meir Golane, c/o Wolfgang Warmbier, Otto-Hahn-Strabe 22, D-7700 Singen, Hohentwiel, West Germany. Tel: (+49) 7731-62061.

## March

**9-11**

**10th International Zurich Symposium & Technical Exhibition on Electromagnetic Compatibility**
Zurich, Switzerland
Information: Symposium Chairman, Dr. Gabriel Meyer, ETH Zentrum-IKT, CH-8092 Zurich, Switzerland. Tel: (411) 256 27 90. Fax: (411) 262 09 43.

**15-18**

**IEEE Multi-Chip Module Conference**
Santa Cruz, CA
Information: MCMC-93, Attn: Jean McKnight, Computer Engineering, University of California, Santa Cruz, CA 95064. Tel: (408) 459-2303. Fax: (408) 459-4829.

**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, UK. 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, UK. Tel: (44) 071 240 1871. Fax: (44) 071 497 3633.

**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.

**18-22**

**The NAB Multimedia World Conference and Exhibition**
Las Vegas, NV
Information: NAB, 1771 N Street, NW, Washington, DC 20036-2891. Tel: (202) 429-5350. Fax: (301) 216-1847.

**28-30**

**EMC/ESD International 1993**
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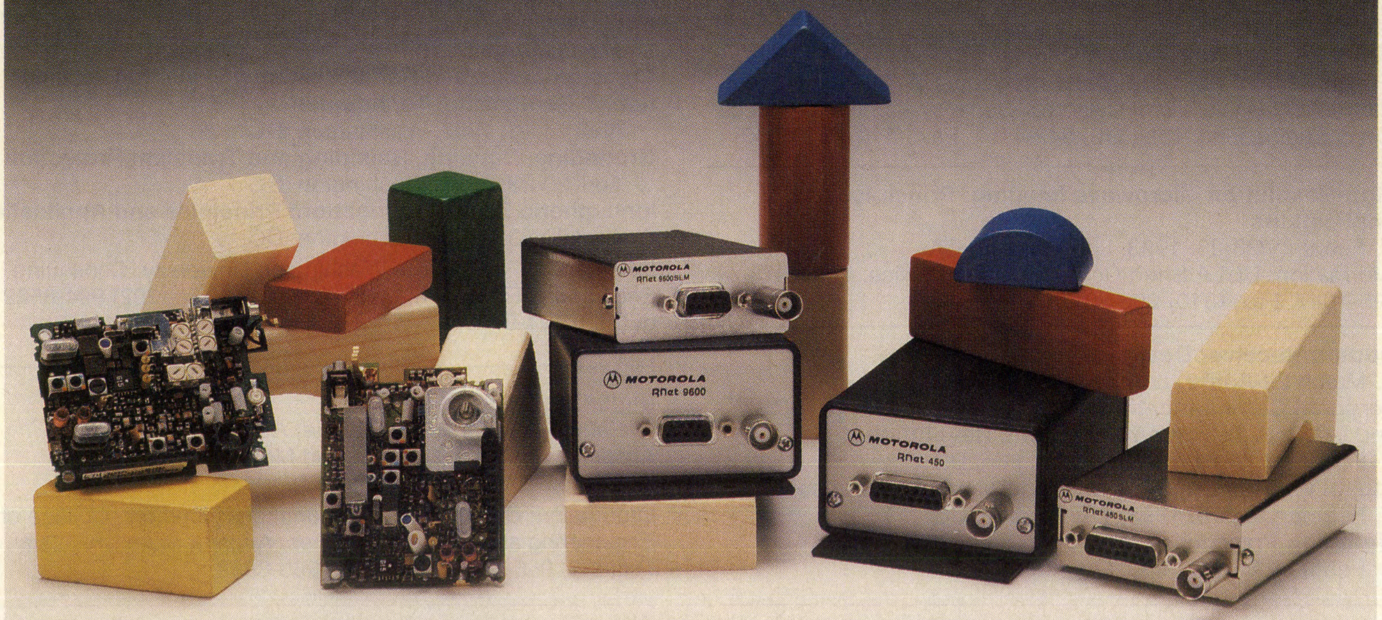


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**Microwave Radio Systems**
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**Modern Digital Signal Processing: Analysis, Design and Applications**
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**Ionospheric Radio Propagation: Principles and Application**
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## Field Strength Comparison Status Updated

In 1991, NIST initiated an international comparison of measurements of electric field strengths for the International Bureau of Weights and Measures. To date, a comparison has been completed with two government and several industrial laboratories in Japan. In 1993 and 1994, NIST plans to perform comparisons with the United Kingdom, Germany, France, Italy, Poland, Austria and Korea. Switzerland, Russia and China also are expected to participate at a later date. The entire project, which should take another four or five years to complete, is needed to show traceability among national standards laboratories. This traceability will become important in 1995 when the 12-nation European Economic Community will require imports to meet standards for electromagnetic compatibility.

**Plan for Testing Semiconductor Packaging** — The Joint Electron Device Engineering Council's JC-15 Committee recently adopted a plan for testing the various modeling and measurement techniques currently in use for thermal and electrical characterization of semiconductor packages. After the participating companies have conducted a round-robin series of tests using their individual methodologies and instrumentation, the JC-15 Committee will analyze the results with the objective of creating industry standards for both modeling and measurement. The JC-15 Committee was established in 1991 as a broad-based industry effort to address the growing need for standard methods to both model a proposed circuit's electrical and thermal packaging requirements and to rate and measure the characteristics provided by various packages.

**Certification Plan for Antenna Performance** — Companies and agencies planning to construct near-field ranges to measure the performance of phased-array antennas will be interested in a new NIST publication outlining a certification plan that tests such facilities for various parameters. Titled "A Certification Plan for a Planar Near-Field Range Used for High-Performance Phased Array Testing" (NISTIR 3991), the document discusses policy issues, measurement requirements and various tests required to characterize errors associated with measurements. Tests include those for alignment error, errors caused by instrumentation, and errors caused by radio-frequency energy traveling undesired paths. NISTIR 3991 is available from the National Technical Information Service, Springfield, VA 22161, (703) 487-4650 for $17 (print) or $9 (microfiche). Order by PB 92-213305.

**New Noise Standards Available** — Three new thermal noise measurement services are being offered by NIST that can benefit manufacturers and users of microwave equipment. The first of these services measures noise standards at discrete frequencies of 12.4, 13.5, 14, 15, 16, 16.5, 17 and 18 GHz using type N and 3.5 millimeter connectors in coaxial cable. The second service measures noise standards from 18 to 26 GHz in

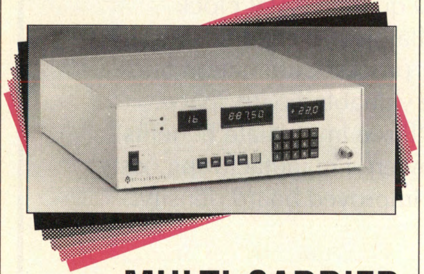
WR42 waveguide, while the third measures noise from 18 to 26 GHz using 3.5 millimeter connectors in coaxial cable. There are specific requirements for reflection coefficient, temperature and excess noise ratio. For additional details on these new services, contact J. Wayde Allen, Div. 813.01, NIST, Boulder, CO 80303-3328. Tel: (303) 497-5871.

**ITU to Restructure** — In order to keep pace with a rapidly changing world telecommunications environment, the International Telecommunications Union recently held an additional plenipotentiary conference to reorganize their internal structure. It is the third major restructuring of the ITU in its 127-year long history. The ITU development machinery, revamped in 1990, will remain unchanged, however, the High Level Committee recommended that the standards setting activities of the CCITT and CCIR be consolidated, thus forming a telecommunications standardization sector. The rest of the CCIR activities, essentially those tasks related to the efficient management of the RF spectrum in terrestrial and space radiocommunications, would be integrated into a new radiocommunications sector along with the activities of the IFRB.

**National Engineer's Week** — February 14-20 has been declared National Engineer's Week and this year is dedicated to the theme "Engineers: Turning Ideas into Reality." An estimated 30,000 engineers will conduct programs emphasizing "Engineering Energy" in elementary, junior and senior high schools across the country. National Engineers Week '93 is jointly sponsored by 18 engineering societies and 10 major corporations, with the cooperation of hundreds of businesses, colleges and government agencies.

**Smart Highway Alliance** — An agreement was recently announced between Texas Instruments and MFS Network Technologies for development of the next generation of Automatic Vehicle Identification and Intelligent Vehicle Highway Systems technologies. The TI/MFS Network Technologies alliance will initially focus on the Electronic Toll and Traffic Management market in the United States. Several such systems have been authorized for completion in the 1990s. Announcements from TI and MFS Network Technologies related to several of these projects are expected in 1993.

**Andrew Awarded Cellular Contract** — Andrew Corporation recently announced that it has signed an agreement with Hongkong Telecom CSL (HKTCSL) to provide advanced digital cellular technology to the underground system of Hong Kong's Mass Transit Railway Corporation (MTRC). The Andrew contract is the result of an agreement between Hongkong Telecom CSL and MTRC to extend cellular service to MTRC passengers. Specifically,







customers to HKTCSL Mobile's new Global System for Mobile digital phone service will be able to make and receive calls and conduct phone conversations without interruption as they travel throughout the MTRC system. The two million dollar contract calls for Andrew to design and provide its Distributed Communications Systems technology to Hongkong Telecom CSL.

**PCS Trials Launched** — Qualcomm and Telesis Technologies Laboratory have announced the joint testing of a Personal Communications Services (PCS) system in San Diego. The system operates in the 1850-1990 MHz band using Qualcomm's CDMA digital transmission technology. The tests are currently underway and are scheduled to continue through the first quarter of 1993.

**KW Microwave Moves** — KW Microwave has announced relocation to a new facility. Their new address is 1985 Palomar Oaks Way, Carlsbad, CA 92009. Tel: (619) 929-9800. Fax: (619) 929-9899.

**Industry Standards Set** — The U.S. cellular standards group, TIA has released standards for Narrow-band Analog Mobile Phone Service. The narrow-band standards, IS-88, IS-89 and IS-90, are based upon Motorola's NAMPS technology.

**Flam & Russell to Upgrade Army Radar** — Flam & Russell, Inc. has been contracted by the U.S. Army Strategic Defense Command to upgrade the AN/FPQ-19 tracking radar at the U.S. Army Kwajalein Missile Range. The upgrade will enable the operators to switch between beacon frequencies in mid-track, so that multiple beacons whose frequencies lie anywhere within the radar operational bandwidth can be tracked sequentially. Other terms of the contract were not released.

**Philips Semiconductors Restructuring** — Philips Semiconductors recently announced the restructuring of its North American integrated circuit operation as part of an ongoing plan to integrate geographically dispersed units into a consistent worldwide organization focused on regional markets. Signetics company will be phased out in its present form, its operations carried on in three organizations that report directly to Philips Semiconductors in The Netherlands: the North American Regional Sales Organization (RSO), the Standard IC Business Group and the Application Specific IC Business Group. The Signetics sales and support infrastructure will be retained within the North American RSO. In addition to North America, RSOs have been established for Europe, Japan/Korea and Southeast Asia.

**BCP Relocates** — Broadband Communications Products has announced its move to a new location. Their address is now 305 East Drive, Suite A, Melbourne, FL 32904. The telephone and facsimile numbers remain unchanged.

**Circuit Components Enters Joint Venture** — Circuit Components, formerly a division of Rogers Corporation, and Micro Substrates, Inc. have formed a joint venture called Micro Substrates Corporation. The joint venture will enable the corporation to manufacture patented ceramic substrates with tungsten-copper vias using CCI's existing manufacturing capabilities. The company's Via Plane™ substrate will be marketed through CCI's sales channels which will be expanded to include the microwave industry.

**Partnership Formed by SPS Technologies and National Magnetics** — SPS Technologies and National Magnetics Corporation have agreed in principle to the formation of a joint venture partnership between SPS Technologies' subsidiary, The Arnold Engineering Co. and National Magnetics to produce soft-magnetic wound core products. The joint venture company, which will be known as National-Arnold Magnetics Company, will satisfy the obligations under all existing Arnold and National production contracts.

**Barnes Engineering Acquires Product Line** — The Barnes Engineering Division of EDO Corporation has announced the purchase of an emission microscopy product, with approximately $4 million of product sales, from KLA Instruments Corporation. The new product line will enhance EDO Barnes' existing range of specialized test instruments for the microelectronic and semiconductor industries and will be combined with Barnes' existing product offerings into a new Commercial Instruments Group.

# A Historical Look at RF Modulation Methods

By Gary A. Breed
Editor

The process of getting information onto an RF signal is modulation, while the recovery of that information is demodulation. In order for RF to do its job of carrying that information from one place to another, both parts of the process must be done properly. Here is a brief historical summary of modulation, to help gain a better perspective on today's complex modulation methods.

The earliest form of modulation was simply turning the RF signal on and off. Using Morse Code or other codes, sequences created by this on-off keying (OOK) could convey information. Receivers were simple, only needing to detect the presence or absence of a signal. This modulation method is still in use — control devices (garage door openers, etc.) often use OOK to keep circuitry simple. Morse Code is used by amateur radio operators who appreciate its reliability and simplicity. Ship-to-shore radios using Morse Code still exist, but nearly all have been retired in favor of satellite communications.

Voice communications required a form of modulation that could change rapidly, following speech patterns. Amplitude modulation (AM) was the first type developed, varying the strength of an RF signal to exactly match the variations in speech and music. AM has the advantage of easy demodulation; a simple diode detector can recover the audio signal. Crystal sets consisting of little more than a diode and an antenna let hundreds of thousands of people hear their first radio broadcasts.

Starting in the 1930s, developments began happening rapidly. Major Armstrong developed frequency modulation (FM), which is more immune to interference than AM, easy to modulate and transmit, but more complicated to demodulate. Compared to AM, FM more easily transmits high-frequency information, which has made FM the dominant method for music broadcasting.

World War II brought radar's development and various types of pulse modulation. Ideally, radar signals are transmitted, then instantly shut off to hear the echoes returning from the target. Actually, the first radar systems used separate transmitting and receiving sites because proper pulse modulation couldn't yet be accomplished. As radar developed, pulse modulation was enhanced by techniques which increase range and accuracy, while making jamming more difficult. Chirped signals, which make a quick sweep across a frequency range, were developed, along with various mathematical rearrangements of pulses into unique wave shapes. These signals required special signal processing using delay lines and filters, which has led to today's digital signal processing.

After the war came television, and a combination of wide bandwidth AM (for the video) and FM (for the sound) was adopted for its modulation and demodulation. Around this time, single-sideband (SSB) communications was developed as engineers found ways to eliminate redundant portions of an AM signal to make a more efficient means of voice transmission.

Along the way, frequency shift keying (FSK) was developed, a fast "wobble" between two slightly different frequencies. This was used for the first transmission of digital data: Baudot codes for radioteletype. FSK is still used for data transmission, using a radio link the same way a facsimile machine or computer modem uses a telephone line.

There are many other milestones in the development of modulation methods, but let's bring the story up to date.

## Modulation Today

All of the exciting new RF applications rely on modulation and demodulation that is highly complex. The purpose of these complicated modulation schemes is to get more information into the limited space of the RF spectrum. In the early days, there was little competition for spectrum space, so simple, brute-force modulation methods worked well. But with the amount of information we wish to communicate to one another, technology must be used to greater advantage. Advanced methods of modulation make it possible to transmit that information without using up all of the available RF frequencies.

These new RF systems involve transmission of information in digital form. The audio compact disc (CD) contains digitized music, and work is underway to replace FM broadcasting with digital audio broadcasting (DAB) to match the CD's level of sound quality. HDTV/ATV will be broadcast as digital information, as well, with new TV sets including extensive digital circuitry in addition to the RF portions. All of the next generation of voice communications (PCN, DECT, digital cellular) send information in digital form. To send this data, two major types of modulation are being used: complex phase and amplitude modulation, and spread spectrum. The first of these is basically a refinement of classic modulation methods, but spread spectrum is a bit different.

Spread spectrum (SS) is a developing modulation technique which makes a trade between amplitude and bandwidth. An unmodulated RF signal occupies just a single frequency, while SS either "smears" the signal across a range of frequencies, or hops around among a number of different frequencies. The advantage of this method is that the same total energy carries the signal, but at any single frequency, the energy is far less than conventional RF signals. This reduces the potential for interference. The precision of SS and other complex modulation is literally mathematical — every type of modulation can be represented by a set of equations. The functions described in these equations establish exactly what the modulator should do regarding amplitude, frequency and phase as it transfers information to an RF signal. The demodulator must operate on the same set of equations in the reverse order, to exactly recover the transmitted data.

To implement the modulation techniques required today and tomorrow, the RF engineer's challenge is to develop circuitry that will maintain the required precision, stability and reliability. *RF*

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# A Low Sideband-Jitter Phase Lock Loop

By Rand H. Hulsing II
Sundstrand Data Control

*This article describes a low frequency implementation of a low sideband-jitter PLL which can be built using readily available, military grade parts for less than five dollars. The improvement over the standard 20 Hz bandwidth multiple frequency lock loop shown in Figure 1 is anywhere from –20 dB to –60 dB. In addition, this improvement tracks over wide temperature and power supply variations. Best results are obtained by using two SE565 PLL chips from the same lot and mounted in close proximity. Further improvement could be obtained with an integrator between the detector and the VCO. This would relieve initial setting of the VCO center frequency and track out jitter over the full capture range of the loop.*

The basic multiplying frequency loop shown in Figure 1 consists of a 400 Hz 50-50 duty cycle output locked to a 100 Hz input. The demodulator produces a typical 2 Vp-p 200 Hz square wave. This is filtered by the two 3.6k resistors and an external 3 uF capacitor, producing a 30 Hz pole for a 20 Hz bandwidth PLL with a damping of 0.5. The error signal present at pin 7 is a 380 mV p-p triangle waveform and is applied to the VCO. This produces the –20 dB sideband jitter shown in Figure 2.

An inverted triangular wave can be generated by adding a second demodulator running only from divisions of the output frequency, but with inverted phasing. This is directly summed with the first demodulator's output by tying the two outputs together (Figure 3). The resulting cancellation produces the clean square wave spectrum shown in Figure 4. The sidebands are –62.5 dB from the 400 Hz fundamental and the usual 1/3 at 1200 Hz, 1/5 at 2k Hz, etc. The circuit divides the 800 Hz VCO down to 400, 200 and 100 Hz using standard MC54HC74's to get 50-50 duty cycle outputs. Note, the second demodulator uses the inverted 100 Hz and a 90 degree phase shifted 100 Hz for its inputs. The closed loop bandwidth is the same 20 Hz second order response with a damping of 0.5.

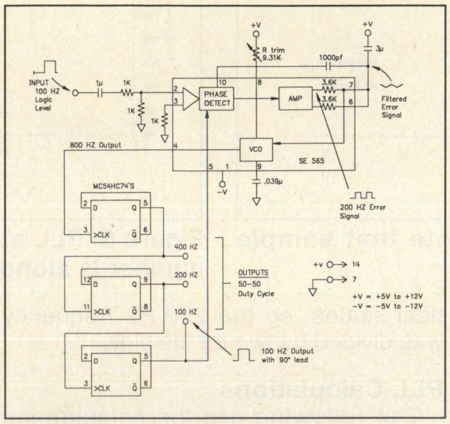

Figure 1. A standard multiple frequency lock loop.

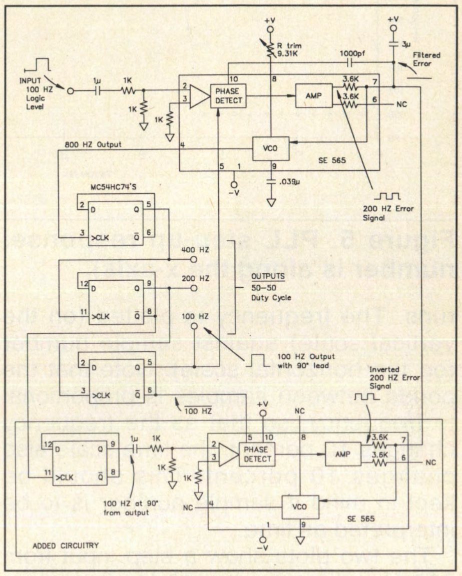

Figure 3. Improved PLL.

The PLL's step response was measured using an HP3571A, which makes up to 1000 continuous period measurements. The counter has two inputs; however, triggering problems allowed only one channel of measurement at a time. Because of this, input and output signals were measured in separate

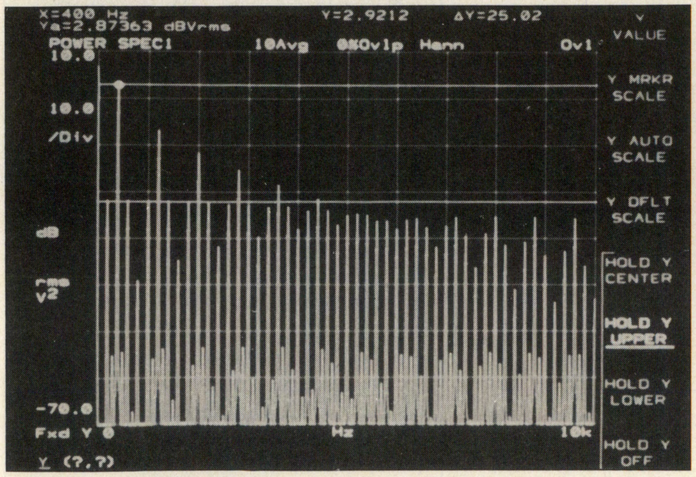


Figure 2. Standard PLL jitter

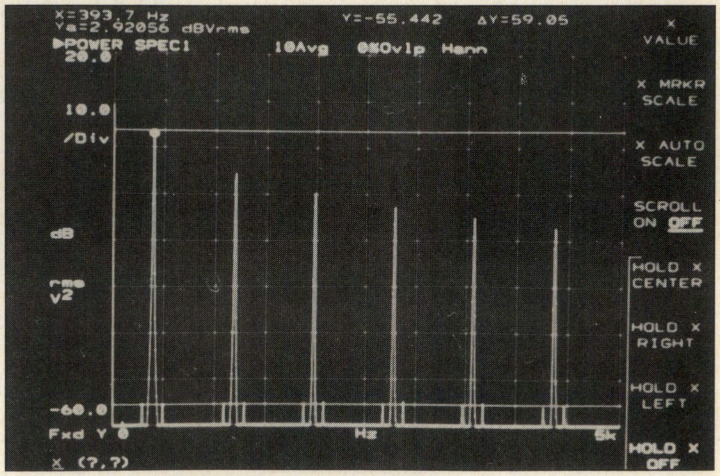


Figure 4. Low jitter PLL spectrum.

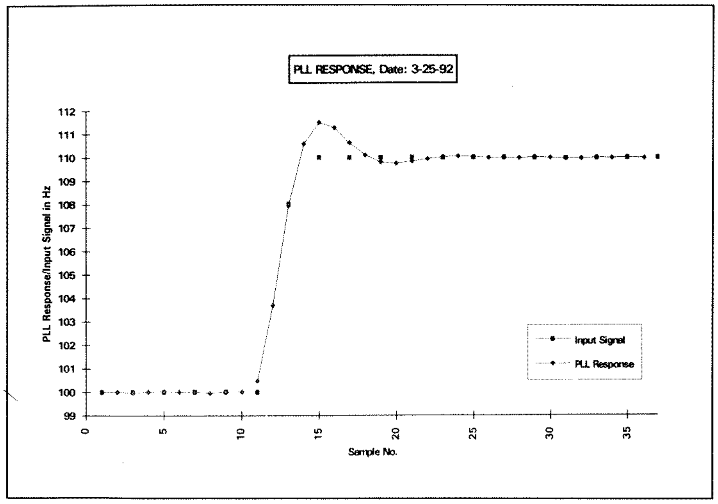


**Figure 5. PLL step-up response, (Note that sample number is along the x-axis).**

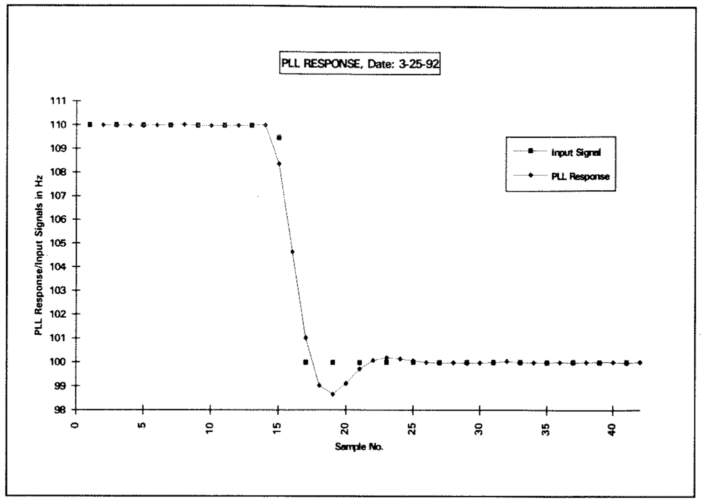


**Figure 6. PLL step-down response, (Note that sample number is along the x-axis).**

runs. The frequency is plotted (on the vertical scale) against sample number (on the horizontal scale). Note that the period between samples is proportional to frequency, so that as the frequency changes 10 percent, the time scale also changes 10 percent. This should be kept in mind if sample number is to be interpreted as time.

The two plots show a step input from 100 Hz, changing to 110 Hz (boxes on plot). The step response was plotted with diamonds, which were lined up with the input (the counter data was asynchronous with the square wave input timing). The response in both directions shows a 0.5 damping characteristic of 15 percent overshoot. The output crosses the input level for the first time after three steps, which would be the response for a 20 Hz second order system. As can be seen, the response is stable and almost textbook in nature. From the previously supplied noise data, this circuit works very well.

Note, the 200 Hz signal was used for output data in order to show the finer details of the step. The data was plotted using Excel which did not allow two vertical scales, so the 200 Hz frequency was divided by two for the plots.

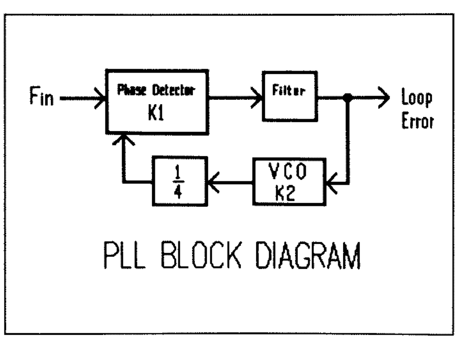


**Figure 7. PLL block diagram.**

## PLL Calculations

The following are the calculations used to obtain the theoretical frequency response of the phase locked loop.

Phase detector gain (Volts/radian):

$$K1 = \frac{2}{\pi} 0.45 = 0.286 \qquad (1)$$

Filter pole (Hz):

1.8k – 3µF

$$WP = 2\pi 29.5 = 185.354 \qquad (2)$$

VCO gain ( (radian/sec)/Volt ):

$$K2 = 2\pi \frac{400}{1.9} = 1.323 \times 10^3 \qquad (3)$$

Open loop gain with divide by four countdown:

$$K = \frac{K1 \cdot K2}{4} = 94.737 \qquad (4)$$

The transfer function includes a sample frequency at double the input frequency. This is included as the sinc function.

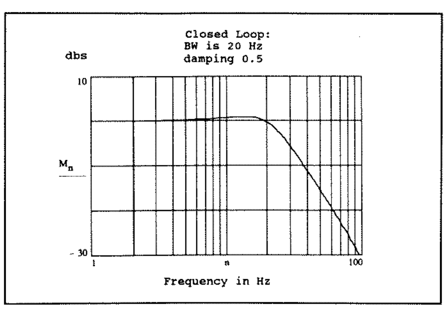


**Figure 8. Bode plot of PLL closed loop transfer function.**

$n = 1 \ldots 100$, $s_n = jn$

FS = 200 (sample frequency in Hz)
FP = 30 (filter pole frequency in Hz)

Inverted open loop gain:

$$G_n = \left(2\pi \frac{s_n}{K}\right)\left(\frac{s_n}{FP} + 1\right) e^{\frac{\pi}{FS} \cdot s_n} \qquad (5)$$

Closed loop expression:

$$C_n = \frac{\sin\left(\pi \frac{n}{FS}\right)}{(G_n + 1)\left(\pi \frac{n}{FS}\right)} \qquad (6)$$

Magnitude of $C_n$ (dB):

$$M_n = 20 \log \left(|C_n|\right) \qquad (7)$$

*RF*

### About the Author

Rand Hulsing is a Research Specialist in Instrument Systems at Sundstrand Data Control, a division of Sundstrand Aerospace. He has been in their Engineering Department for 15 years. He has worked on design and development of accelerometers, inclinometers, magnetometers, oil well survey tools and is currently developing an accelerometer/rate sensor in a single package. He has 24 patents issued and several pending. He can be reached at Sundstrand Data Control, 15001 N.E. 36th St., M/S 21, P.O. Box 97001, Redmond, WA 98073-9701, or by phone at (206) 885-8575.

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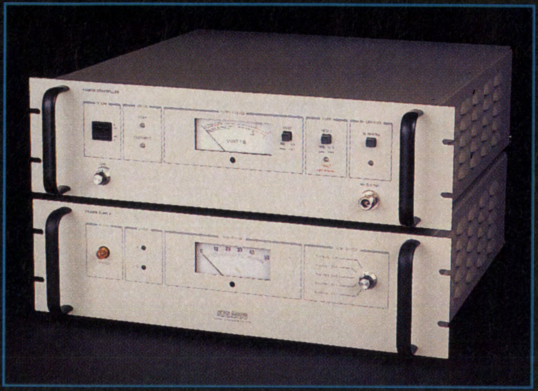

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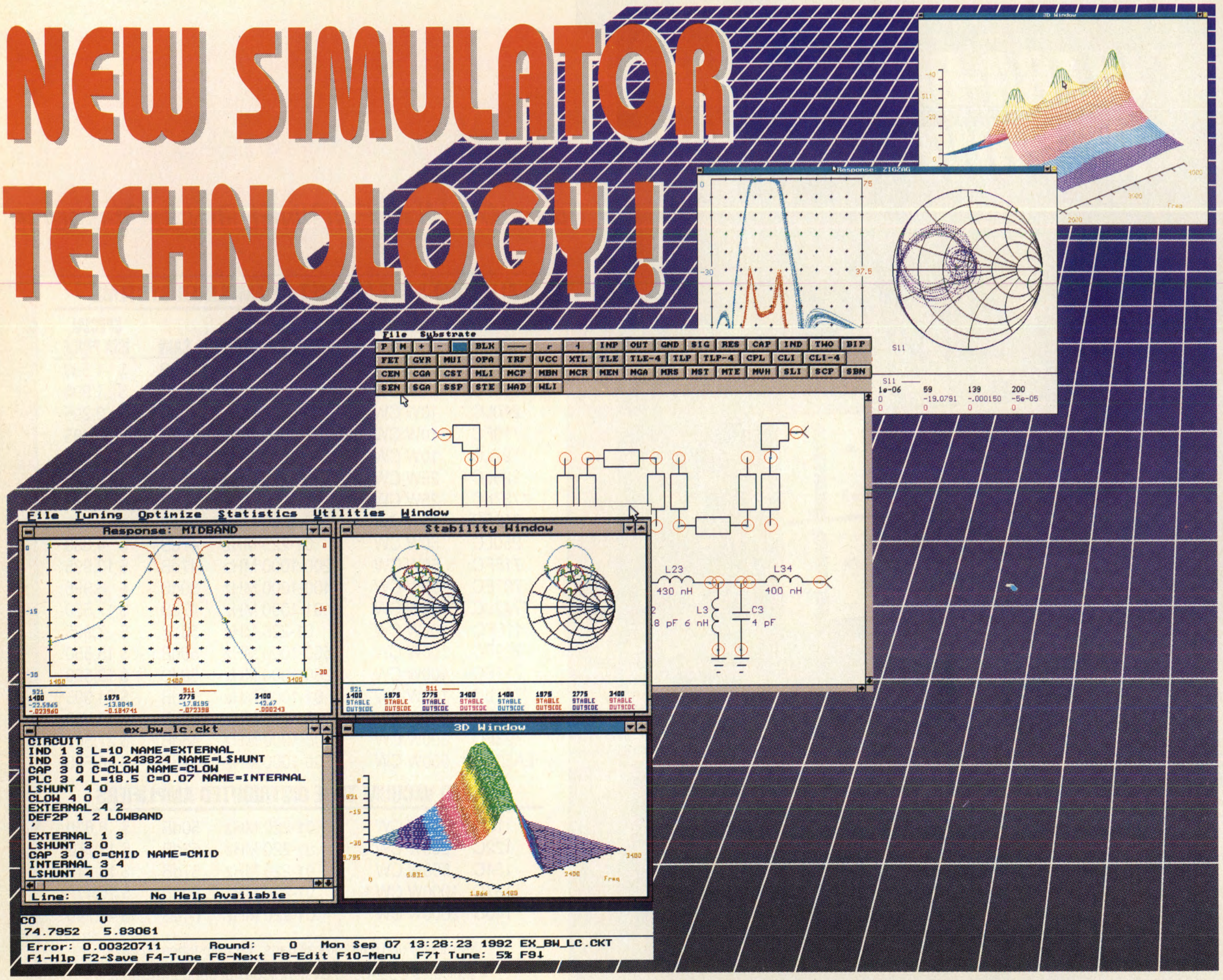


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# 1.8 GHz Direct Frequency VCO With CAD Assessment

*By Brendan Kelly, Dr. Noel Evans, Brian Burns*
*University of Ulster at Jordanstown*

*The ever increasing consumer demand for the radio frequency spectrum in personal and mobile communication systems is driving RF design into the GHz bands. Computer simulation of RF circuits is helping to resolve the inherent design problems as software packages such as EEsof's jOMEGA become available. How accurately a simulated RF circuit performs in practice is still, however, a matter of some conjecture. This paper describes a novel VCO design, in concept and practice, and compares its actual and simulated performance.*

RF source design at frequencies above 1 GHz still relies to a large degree on the discrete component construction of an oscillator followed by cascaded mixer and multiplier circuits. Direct frequency generation offers:
1) A reduced component count resulting in space and power efficiency and,
2) The elimination of close-in spurious outputs and hence no intermodulation problems.

Cavity and line resonators may be used to control direct frequency generation, but cavities are physically bulky and line resonators have low Q-factors. Circuits to implement a stable 1.8 GHz oscillator within a miniature package and with low output phase noise present a design problem which can be partially solved by the introduction of high Q-factor coaxial ceramic resonators as the frequency determining elements.

## Resonator Characteristics

Electrically, this ceramic component forms a coaxial conductor which is short-circuited by its silvered outer coating: a typical resonator designed for 1.8 GHz operation has the dimensions shown in Figure 1a. Its equivalent circuit corresponds to that of the parallel RLC circuit shown in Figure 1b (1). Values for $C_p$ and $L_p$ may be calculated from equations 1 and 2, derived from coaxial transmission line parameters (2).

$$C_p = \frac{2\pi\varepsilon_0\varepsilon_r l}{2\ln\left(\frac{D}{d}\right)} \qquad (1)$$

$$L_p = 8\frac{\mu_0\mu_r}{2\pi^3}\ln\left(\frac{D}{d}\right)l \qquad (2)$$

These formulae, together with the Q factor (615) as measured by the manufacturer (1), produce circuit values of:

$C_p$= 7.26 pF $L_p$= 0.854 nH $R_p$= 7.9 kohms

## Negative Resistance Approach to Oscillator Design

Consider an ideal series tuned circuit (infinite Q). When excited, it will oscillate indefinitely because there is no resistance to dissipate energy. In a circuit with a finite Q, oscillations will decay as

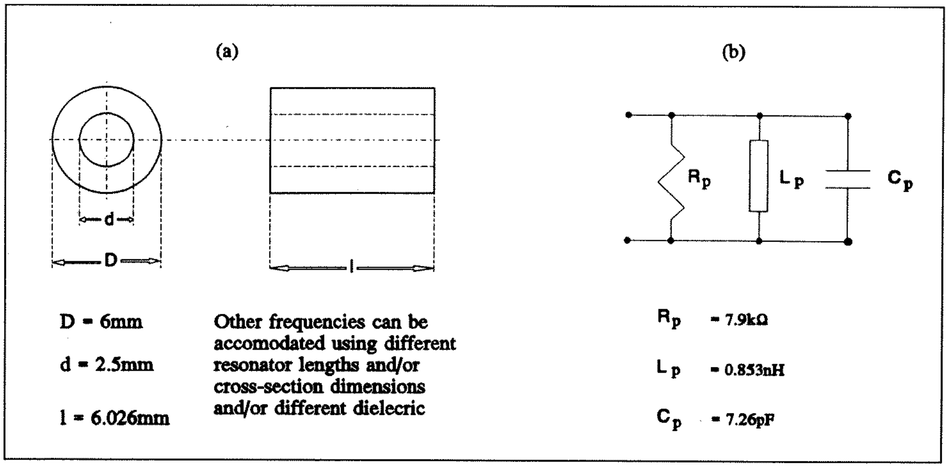



**Figure 1. 1.8 GHz resonator and equivalent circuit.**

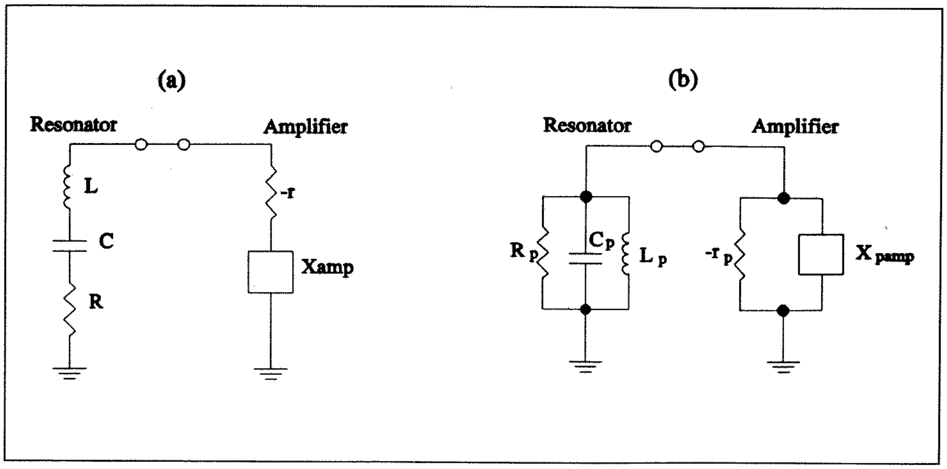



**Figure 2. Negative resistance and conductance.**

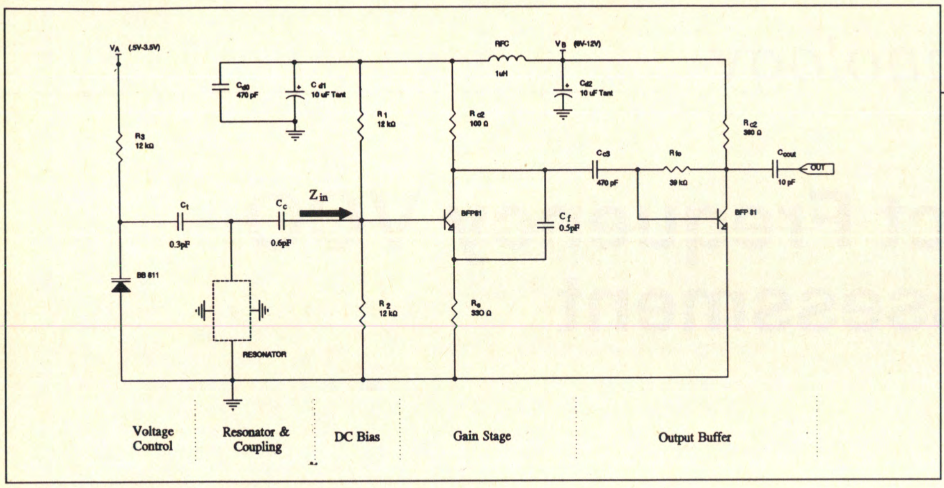



**Figure 3. Wideband oscillator circuit.**

energy is dissipated. The function of the amplifier in a practical oscillator circuit is to overcome any loss resistance by supplying an amount of energy equal to that dissipated, thereby sustaining oscillation. This energy source may be interpreted as a negative resistance (–r) in series with the tuned circuit, as shown in Figure 2a.

Alternatively a series circuit can be transformed to a parallel circuit to become a negative conductance category oscillator, as shown in Figure 2b. In this form the resonator in Figure 2b can be more easily related to the coaxial ceramic resonator equivalent circuit and the capacitive pulling effect of the amplifier reactance on the oscillating frequency can be envisaged.

It will be shown, using both traditional scattering/admittance/impedance techniques and EEsof's simulation software, that the amplifier chosen for this work has a negative input resistance.

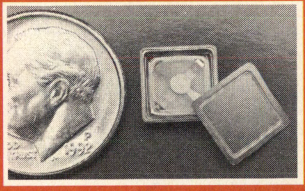
## Conditions for Oscillator Start-Up — Traditional Method Vs. Simulated

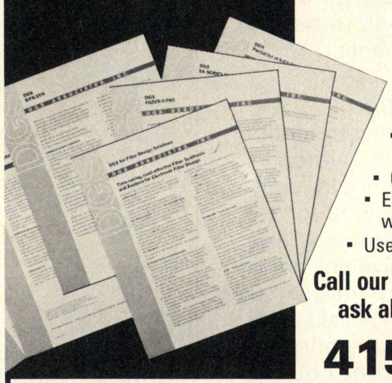
The circuit of Figure 3 was implemented in surface mount form. To establish that conditions for oscillation exist, it is necessary to show that the real part of its input impedance, $Z_{in}$, is negative.

The gain stage of Figure 3 may be built around a three-terminal device as shown in Figure 4a, where N is the two-port representation of Q1 (3, 4). $Y_c$ represents the parallel admittance of the feedback capacitor $C_f$ and $Y_r$ the series admittance of $R_e$. Admittances $Y_b$ and $Y_l$ represent the parallel combination of the two 12 kohm bias resistors and the load, respectively. $Y_b$ is insignificant compared with the other parameters and may be neglected.

This three-terminal device can be represented by the parallel and series combination of two-port networks as shown in Figure 4b.

The transistor, capacitor and resistor networks may be defined in terms of their short-circuit admittance (Y) para-



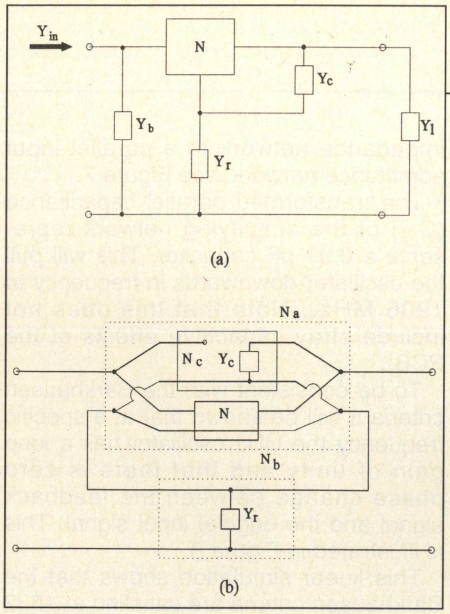



**Figure 4. a) 3-terminal representation of Q1 with parallel $C_f$ and series $R_e$, b) parallel and series combination of three two-ports.**

meters. Y-parameters for the transistor network can be derived from the data sheet S-parameters to give the following admittance matrix:

$$Y_T = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} = \tag{3}$$

$$\begin{bmatrix} -0.5967 + 0.2423i & 0.0762 + 0.0420i \\ 1.4072 + 2.0784i & 0.1897 - 0.3246i \end{bmatrix}$$

The Y-matrix for the capacitor as a two-port network is:

$$N_c = \begin{bmatrix} 0 & 0 \\ 0 & Y_c \end{bmatrix} \tag{4}$$

The Y-matrix for the resistor as a two-port network is:

$$N_b = \begin{bmatrix} Y_c & Y_c \\ Y_c & Y_c \end{bmatrix} = Y_c \begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix} \tag{5}$$

The Y-matrix for the complete network as illustrated in Figure 5 is:

$$Z_N = \left[ \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & Y_c \end{bmatrix} \right]^{-1} + \frac{1}{Y_r} \begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix} \tag{6}$$

where

$$Y_c = 2\pi f C \ \ S, \quad Y_r = 330 \ \ S \tag{7}$$

and f = 1.8x10$^9$ Hz.

This matrix may be reduced to the normal 2 X 2 form. Then, the input impedance of the gain circuit can be shown to be:

$$Z_{in} = Z_{11} - \left( \frac{Z_{12} Z_{21}}{Z_l + Z_{22}} \right) \tag{8}$$

Figure 5a shows how the input impedance in this case varies with values of feedback capacitance, $C_f$. It can be seen from this graph that the network exhibits real negative input impedance, and will, therefore oscillate for $C_f$ values from 0.3 pF to 1.9 pF. The capacitive reactance will act in such a way as to reduce the frequency of oscillation.

The results of Figure 5a can be compared with the jOMEGA linear simulation of Figure 5b, this shows very similar input impedance characteristics for the same values of feedback capacitance, $C_f$.

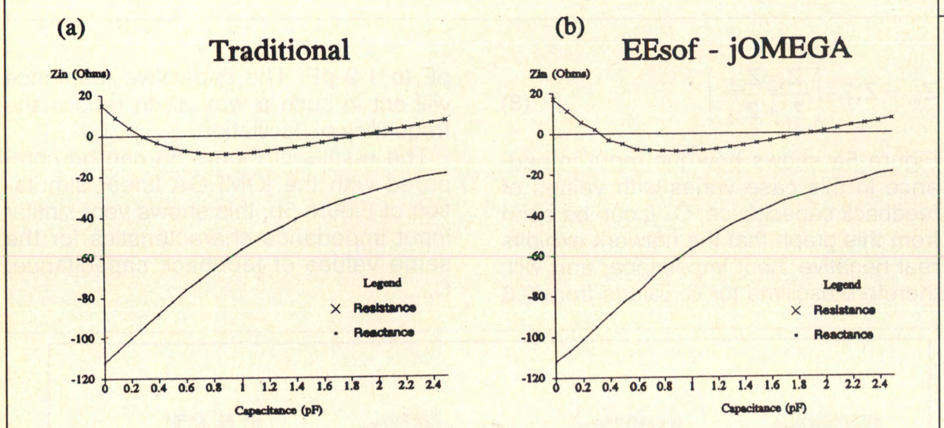



**Figure 5. $Z_{in}$ evaluated by traditional and simulated methods.**

### Frequency of Sustained Oscillation — Theoretical, Simulated and Measured

When the coaxial ceramic resonator is attached to the amplifying network, oscillations will begin if the magnitude of $R_{in}$ is numerically greater than the resistance of the resonator. These conditions are easily seen if both the amplifying network and the resonator are shown as series resistances. This is illustrated in Figure 6.

In an operating circuit the capacitive reactance will act in such a way as to reduce the frequency of oscillation (i.e. as a capacitor in shunt with the resonator). To illustrate this point it is appropriate to transform the series input impedance network to a parallel input admittance network: see Figure 7.

The transformed parallel capacitance ($C_{pa}$) of the amplifying network represents a 0.91 pF capacitor. This will pull the oscillator downwards in frequency to 1906 MHz. (Note that this does not include stray capacitive effects of the PCB).

To be consistent with the Barkhausen criteria it will be shown that at a specific frequency the UUJ oscillator has a loop gain of unity and that there is zero phase change between the feedback signal and the original input signal. This is illustrated in Figure 8.

This linear simulation shows that the Barkhausen criteria are satisfied at 1832 MHz. However, the frequency of oscillation will increase from the linear condition as the nonlinear condition is evaluated (5) and therefore the frequency of oscillation can be expected to be a little higher, perhaps about 1900 MHz.

In practice, when stray capacitances on the oscillator PCB were kept to a minimum, the frequency of oscillation

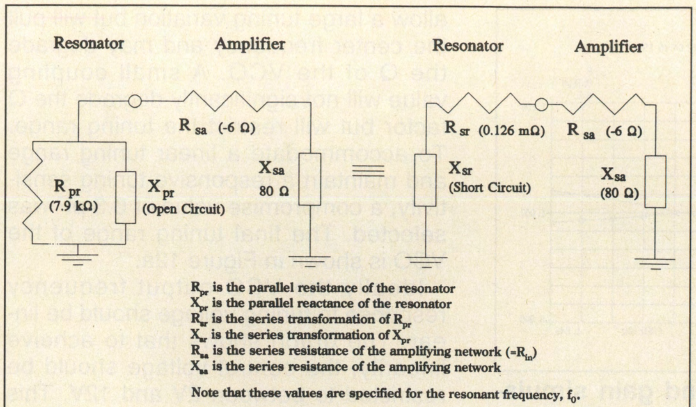



**Figure 6. Series equivalents of resonator and amplifying network to illustrate that oscillations will begin.**

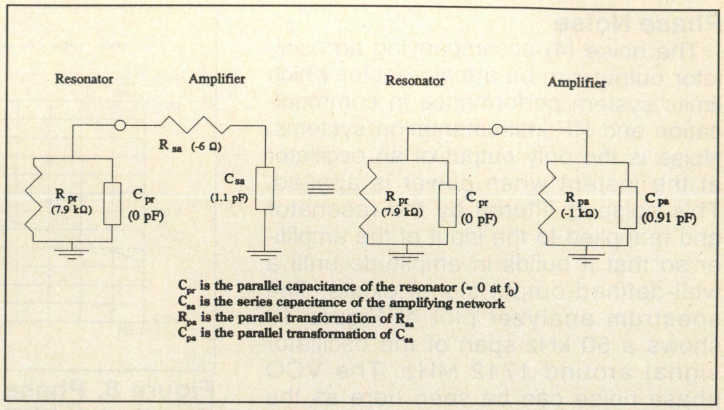



**Figure 7. Parallel equivalents of resonator and amplifying network to illustrate the pulling effect of the shunt capacitance.**

was raised to 1.9 GHz. This is consistent with the theoretical and simulated results introduced above.

### Oscillator Characterization

Figure 9 shows the surface mount VCO (top screen removed for clarity) which has been developed from a prototype built on copper ground plane (G10 glass fiber PCB material). The PCB is of a double layer type designed to increase the heat dissipation potential of the board and provide excellent shielding and grounding capability. The VCO is decoupled using both in-circuit and feedthrough devices. Output power is delivered via an on-board impedance matching buffer and SMA series connector.

In characterizing this VCO, phase noise, tuning, residual FM and pushing will be discussed in addition to comparisons between actual and simulated power output spectra.

## Phase Noise

The noise (4) accompanying an oscillator output can be a major factor which limits system performance in communication and RF instrumentation systems. Noise is the only output of an oscillator at the instant when power is applied. This noise is filtered by the resonator and reapplied to the input of the amplifier so that it builds in amplitude until a well-defined output signal exists. The spectrum analyzer plot of Figure 10 shows a 50 kHz span of the oscillator signal around 1742 MHz. The VCO phase noise can be seen here as the sloping sides above and below the center frequency; its characteristics, as measured, are shown in Figure 11a. These surpass the typical response of a commercially available VCO with which they are compared.

## Residual FM

Residual FM (6) is a measure of short-term instability caused by low frequency signals modulating the carrier and is specified in terms of the p-p deviation in Hz. Figure 11b shows the center frequency deviation, measuring 12 kHz.

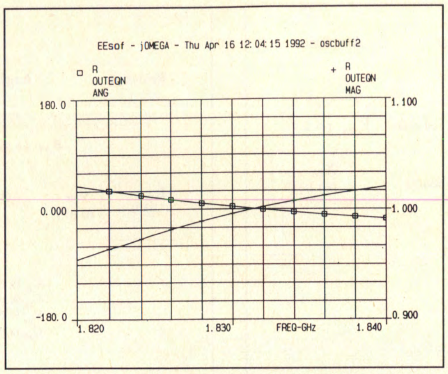



**Figure 8. Phase and gain simulation to establish Barkhausen criteria.**

## Tuning

The tuning range of the VCO depends mainly on the degree of capacitive coupling between the oscillator and the varactor. (See Figure 3 for circuit diagram). A high value of coupling capacitance will allow a large tuning variation but will pull the center frequency and may degrade the Q of the VCO. A small coupling value will not significantly degrade the Q factor but will restrict the tuning range. To accommodate a linear tuning range and maintain a responsive tuning sensitivity, a compromise value of 0.3 pF was selected. The final tuning range of the VCO is shown in Figure 12a.

Ideally the VCO output frequency response to tuning voltage should be linear. The graph shows that to acheive linearity, the control voltage should be restricted to between 2V and 12V. This allows the VCO to be tuned over 40 MHz, giving a sensitivity of 4 MHz/Volt.

## Pushing

Pushing is defined as the change in oscillator output frequency resulting from a change in oscillator supply voltage.

The feedback capacitance, $C_f$, played an important role in this VCO stability factor. $C_f$ was selected for easy oscillator starting, within the negative input resistance range (see Figure 5): an optimum value of 0.5 pF was chosen. Figure 12b shows the pushing characteristics of the VCO for these conditions. The VCO frequency variation is 4 MHz over a supply voltage change from 5V to 12V ie: approximately 570 kHz/Volt.

## Harmonic Suppression

This is defined as the worst-case-amplitude, harmonically-related signal, relative to the fundamental power level, and is measured in dB relative to the carrier (dBc). This measurement is important, for example, when matching the VCO to the prescaler of a PLL.

Harmonic suppression for the UUJ VCO is better than –25 dBc, compared with –15 dBc for a typical high quality VCO. The practical harmonic content, as viewed on a Tektronix 2756P 21 GHz spectrum analyzer, and the simulated output may be compared in Figure 13. The supply voltage was 8V for both.

This figure shows that the difference between fundamental and harmonics is better than –25 dBc. It also demonstrates the accuracy of the simulation software. The simulated fundamental is within 1.5 dB of the measured data and the worst case is a difference of 2.5 dB for the second harmonic. Figure 14 shows the simulated output voltage waveform for a 9V supply.

It can be seen that the output voltage waveform is fairly pure and produces almost 1.4 V p-p into a 50 ohm load,

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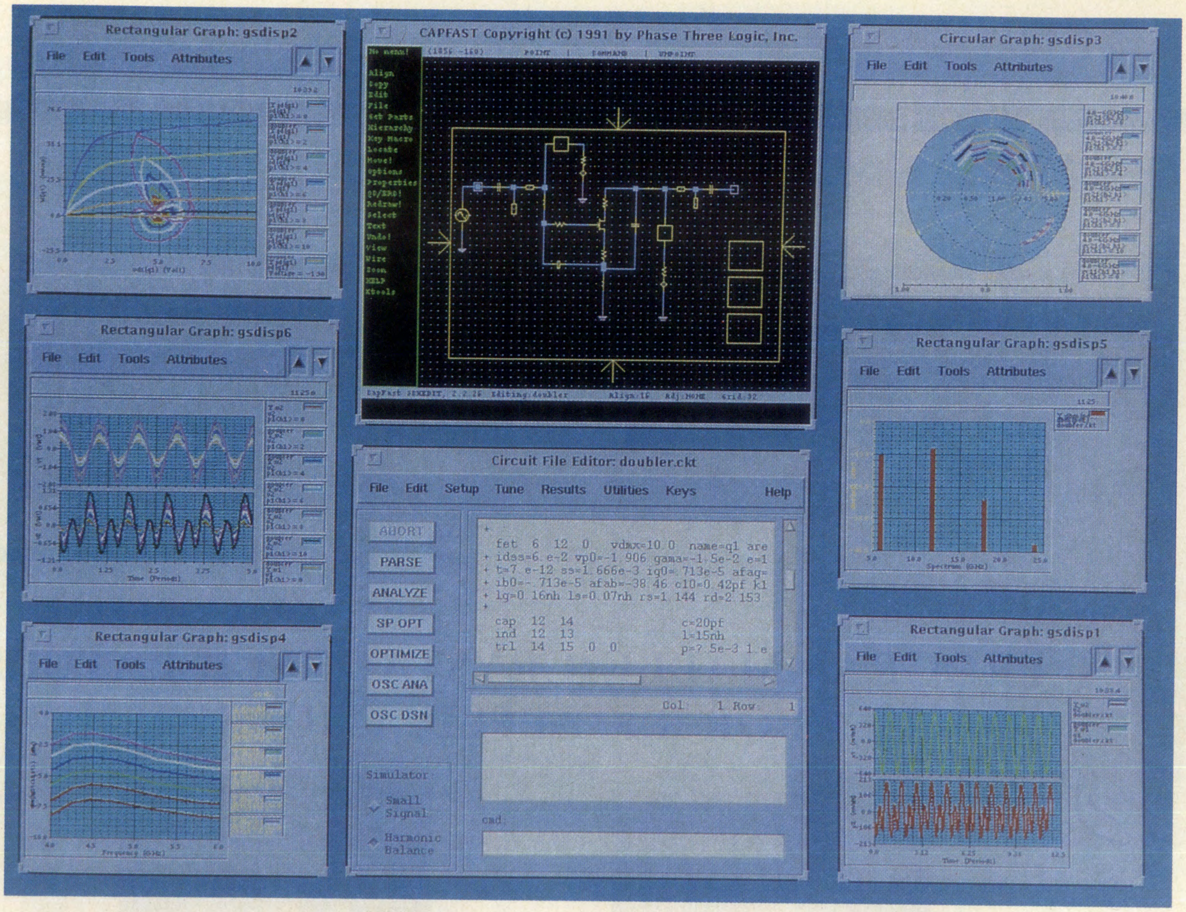


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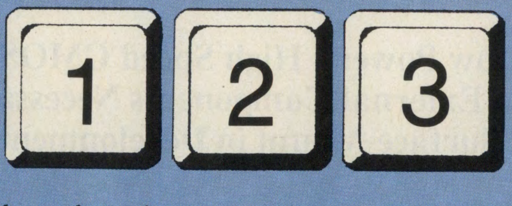


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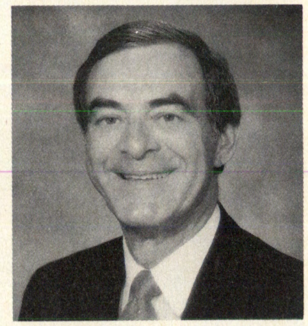

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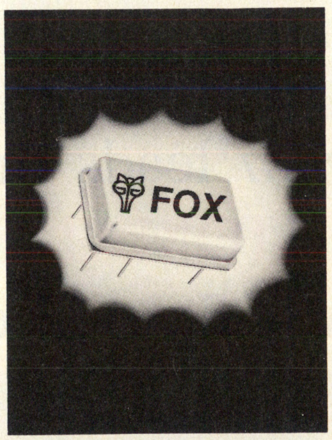


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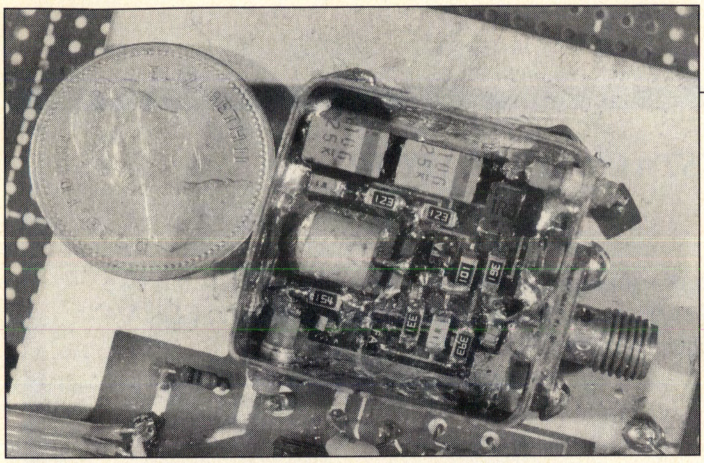

Figure 9. Developed surface mount VCO.

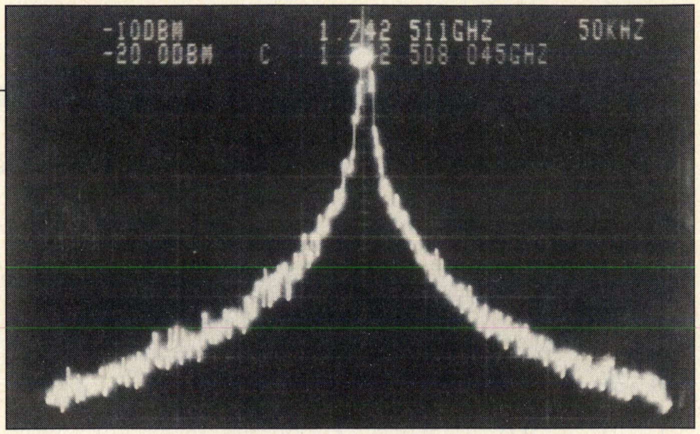


Figure 10. Oscillator spectrum in a 50 kHz span around 1742 MHz.

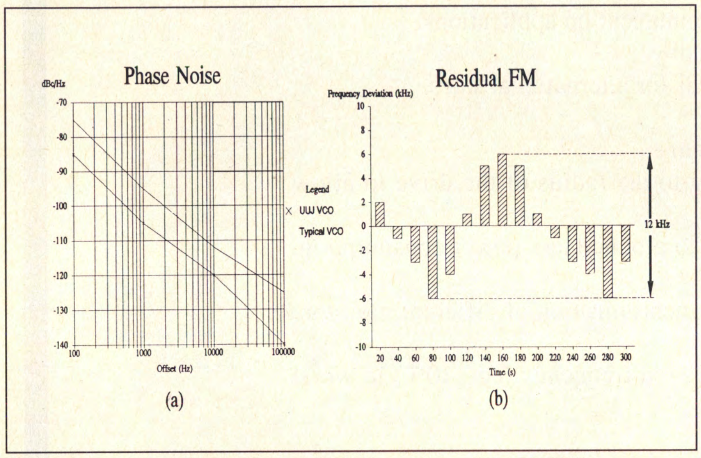


Figure 11. VCO phase noise and residual FM.

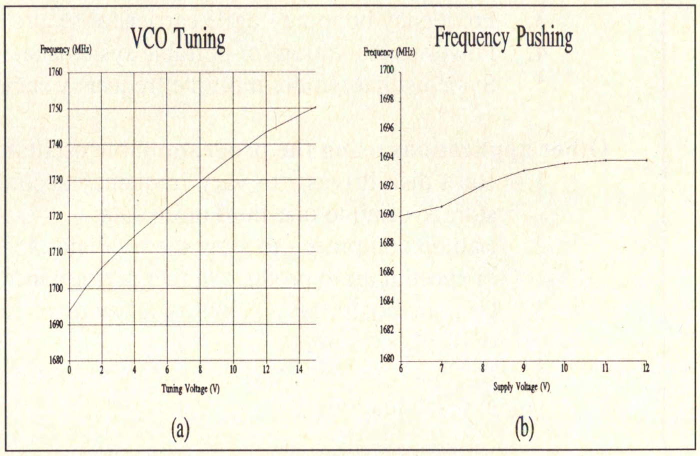


Figure 12. VCO tuning and pushing.

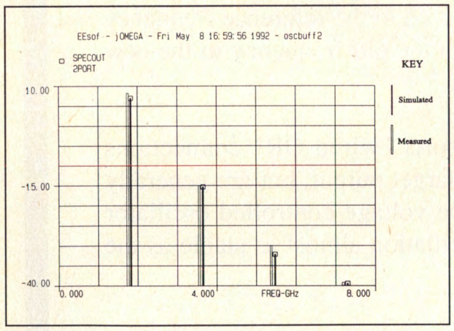


Figure 13. jOMEGA simulated power spectrum in comparison to measured results for similar applied voltages.

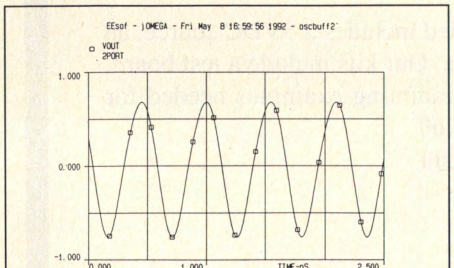


Figure 14. jOMEGA simulated output voltage waveform.

equivalent to an output power of 12.8 dBm. This is very close to the practical result for the output power at 9V: this was 11.8 dBm. The period of the waveform is 0.55 ns, representing a frequency of 1.8 GHz.

## Summary

This paper has described the theoretical and practical design of a high stability VCO for the 1.8 GHz band and compared its performance to that of the same design simulated on EEsof's jOMEGA. It has been shown that the simulation compares favorably with traditional and practical results and goes some way to establishing the viability of jOMEGA for RF circuit design at upper UHF frequencies. *RF*

## Acknowledgements

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# A Flexible Fractional-N Frequency Synthesizer for Digital RF Communications

By Jonathan Stilwell
Philips Semiconductors

*In digital RF communications systems, typical single-loop synthesizers may not meet all switching time, noise and spurious output requirements. An introduction to fractional-N frequency synthesis and a comparative evaluation between standard and fractional-N single-loop synthesizers in a 900 MHz application are presented. A simple passive filter design procedure, information on loop analysis and optimization using commercially available software and design tips are also included.*

First, we will attempt a fast-switching design using a standard PLL. For designs with channel spacings less than 100 kHz, a synthesizer with high frequency resolution is desired. Typically such devices have been used in analog applications with very low loop bandwidths. Such designs meet noise (residual FM, integrated jitter) and spurious sideband requirements but are often very slow to switch channels.

The channel spacings, noise and spurious requirements for some of the emerging digital communications systems are approximately the same as analog systems, but the switching times allowed are much smaller. This poses a challenge for the designer not wanting to design complicated loop filters or revert to a double-loop approach.

Design targets for a narrowband digital RF system local oscillator are shown in Table 1. A PLL frequency synthesizer with standard architecture (Fujitsu MB1501) was chosen and configured with a single loop filter (i.e., no switched loop filter or loop adaptation). As the switching time target is aggressive, the closed loop bandwidth was chosen to be fairly wide (4 kHz) relative to the comparison frequency (30 kHz).

Figures 1 and 2 show the loop filter and resulting close-in noise spectrum. The SSB phase noise at a 1 kHz offset was:

$$L(1\text{ kHz}) = -69.7\text{ dBm} - (-11\text{ dBm}) = -58.7\text{ dBc/Hz}$$

This was too high to meet the specification, as was the 30 kHz spurious at –52 dBc.

The simplest approach to solving such noise and spurious problems would be to reduce the loop bandwidth at the expense of switching time. However, Figure 3 shows that the switching time goal is not being met, either, so the loop bandwidth cannot be reduced. Without a different approach, or a very elaborate loop filter, we are hard pressed to meet the design targets with a single synthesizer.

## Introduction to Fractional-N Synthesis

A typical synthesizer such as the MB1501 above has a frequency resolution or step size equal to the phase detector comparison frequency. With the fractional-N technique, however, the step size is a fraction of the comparison frequency. For example, the design that follows yields 30 kHz resolution from a 240 kHz comparison frequency. It increases the resolution by alternating overall division ratios between two values (for most output frequencies), with the time spent in each ratio under control of logic circuitry. When different channels are selected the times are adjusted so that the overall division ratio changes in fractional (rather than the

| | | |
|---|---|---|
| VCO frequency | $f_{VCO}$ | 918 MHz |
| Channel spacing | $f_{SP}$ | 30 kHz |
| Spurious Sidebands | $\lvert spurs \rvert$ | <–55 dBc |
| SSB Phase Noise | L(f) | <–65 dBc/Hz, 500 to 5000 Hz |
| Switching Time | $t_S$ | <1 ms to within 1 kHz, 15 MHz step<br><2.5 ms to within 100 Hz, 15 MHz step |

**Table 1. Synthesizer target specification.**

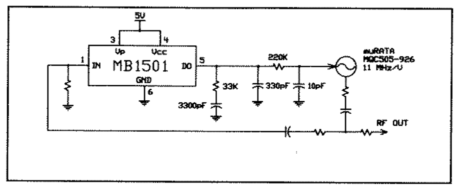



**Figure 1. Standard synthesizer design schematic using MB1501.**

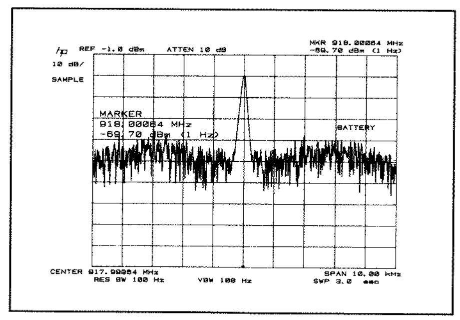



**Figure 2. Close-in noise spectrum of standard design.**

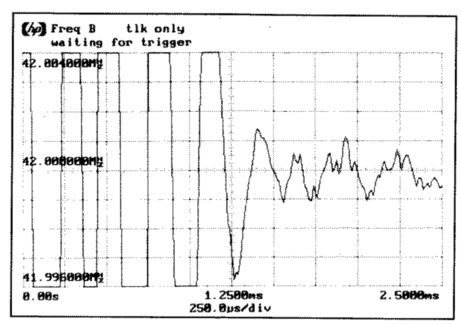



**Figure 3. Switching response of standard design (output signal mixed down to 42 MHz).**

typical integer) increments.

For a given step size this increase in resolution means a higher comparison frequency and therefore a lower overall division ratio. In most synthesizer applications the phase noise is largely dependent on this overall division ratio — the larger the ratio, the higher the noise. With fractional-N division the close-in spot SSB phase noise could theoretically be lowered by up to 20log(240/30), or about 18 dB for our example. Having this lower noise floor and the higher comparison frequency available, a designer who uses the fractional-N method can build circuits with wider loop bandwidths yet maintain the same stability. Such circuits could switch much faster for a given integrated phase jitter, be much quieter for a given switching speed, or be both faster and quieter than conventional designs.

Expensive discrete fractional-N implementations have been used in synthesized signal generators and military equipment for some time. The technique was described by Dr. Ulrich Rohde in *RF Design* (1) and also in his book on frequency synthesis (2). An analysis and description was presented by Roland Hassun of Hewlett-Packard in an article in *Microwaves & RF* (3).

*How fractional-N works* — Figure 4 shows the contents of a digital accumulator used in fractional-N synthesizers. For each channel assigned, the division ratio includes a fractional part that is programmed into the synthesizer as a fractional word (NF). Each comparison cycle the accumulator contents are incremented by the fractional word, leading to overflow when the value exceeds the comparison frequency-to-step size ratio (240/30 = 8 in our example). The output is fed to the dividers so as to swallow a VCO cycle (increase the divide ratio by one) for one comparison cycle each time the accumulator overflows. Thus, over a number of cycles the average division ratio includes a fractional part. It is this fractional part of the division ratio that allows the system to increase the frequency resolution and therefore use a higher comparison frequency.

| Division | Accumulator Contents (NF=3) |
|---|---|
| N | 3 |
| N | 6 |
| N+1 | 1 |
| N | 4 |
| N | 7 |
| N+1 | 2 |
| N | 5 |
| N+1 | 0 |

**If we compute the average division over one complete cycle we get,**

$$N_{ave} = \frac{8N+3}{8}$$

$$= N + \frac{3}{8}$$

**Figure 4. Digital accumulator contents, fractional-N system.**

## An Integrated Fractional-N Synthesizer

The UMA1005 is the first commercially available synthesizer known to have fully integrated fractional-N operation, including accumulators and circuitry for fractional compensation. Figure 5 shows the block diagram of the device. It is designed in the Philips SACMOS process, allowing high frequency operation with very low current (up to 5 MHz comparison frequency, <5 mA typical). Support of external two-, three- and four-modulus prescalers (e.g., the NE701/2/3 family) is provided. The two latter types reduce the minimum required system division ratio below the typical value imposed ($P^2$–P) by standard dual modulus devices. This allows support of the higher comparison frequencies used in fractional-N operation. The choice of a two-chip solution allows for maximum flexibility, as the UMA1005 can be used with various prescalers not only for IS-54 dual-mode cellular and trunked radios but also for 1.9 GHz PCN and 2.45 GHz ISM band applications.

The UMA1005 includes a second, non-fractional-N, PLL frequency synthesizer available that can be used as an offset oscillator for full duplex systems. In a typical system the receive channels might be 45 MHz above the transmit channels. In such a system the main fractional-N synthesizer would be used to synthesize a high-side receive oscillator and the auxiliary operated at 90 MHz. Direct frequency modulation (FSK/GMSK) could be done on the auxiliary VCO. The two signals would be mixed and the difference frequency transmitted.

Use of fractional-N and an offset oscillator improves the phase noise and residual FM of both loops by reducing the overall division ratio compared to standard synthesis of two UHF oscillators. The offset architecture also allows direct VCO modulation down to low frequencies while maintaining fast switching.

Rapid channel changes are achieved due to several device features. First, the device has the high phase detector bandwidth (5 MHz) required for the design of low noise loops with channel spacings from 10 to 5000 kHz. Second, the high speed (10 Mbit/s) three-wire serial interface reduces the time required to initially send channel frequencies from the system microcontroller. Third, partitioning of the data registers allows the output frequency to be changed (after initial loading) by only reprogramming a single 24- or 32-bit word. Fourth, as discussed below, the device reduces switching time by having a high gain mode that allows faster initial acquisition.

The UMA1005 simplifies the design of stable adaptive loops. In such systems the loop bandwidth is increased during switching to allow faster initial frequency acquisition. A unique feature of the UMA1005 is that, unlike methods that switch in various filter components, stable adaptive designs require no increase in the external loop filter component count. This is because the device has dual detector outputs that can be connected to the standard loop filter so as to not only increase the loop bandwidth but also keep the phase margin constant when changing between switching (speed-up) and steady state (normal) modes. Because the UMA1005 integrates high quality charge pumps, most

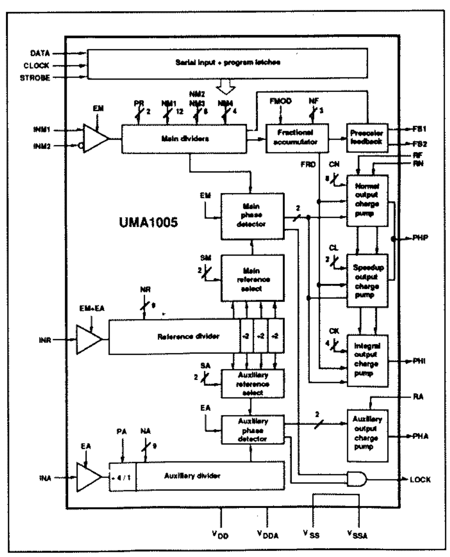



**Figure 5. Block diagram of UMA1005.**

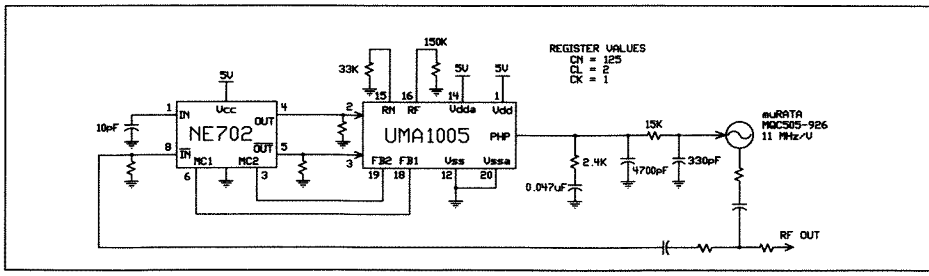



**Figure 6. Nonadaptive UMA1005 design schematic and register values.**

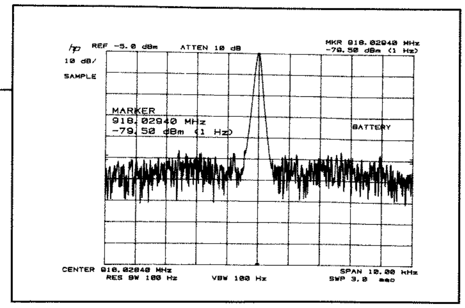



**Figure 7. Close-in noise spectrum of nonadaptive UMA1005 design.**

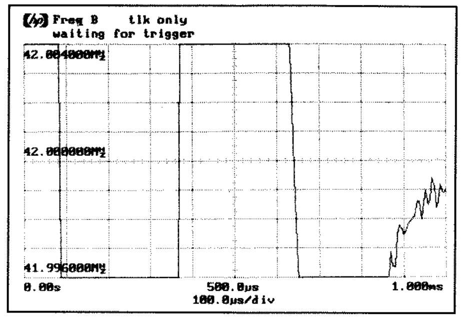



**Figure 8. Switching response of nonadaptive UMA1005 design (output signal mixed down to 42 MHz).**

designs can be realized with a simple loop filter having three to five passive components.

In some existing designs the switchover time of adaptive loops is fixed because the in-lock detector determines which detector or pump is active. With the UMA1005 the adaptation time is user programmable. This allows the user the flexibility to remain in the high gain mode beyond the initial lock detect. This may lower switching time compared to typical designs that may switch to a narrower loop bandwidth before adequate settling has occurred. Waiting to switch modes can also be useful in TDD transceivers that can remain in high-gain mode to make fast RSSI measurements.

*Owning up to the design tradeoff* — The primary challenge of designing a good fractional-N synthesizer is in canceling spurious outputs. When the fractional word is not zero (four-fifths or seven-eighths of the channels with the UMA1005) there will be fractional spurious sidebands in addition to the comparison frequency sidebands. These additional undesired outputs are typically referred to as the fractional spurs and result from fractional jitter. They arise because, although the average division is correct in the system, the instantaneous phase of an uncompensated system is not constant. There is a ramp error in phase each time the accumulator overflows and the overall division value is changed.

Cancellation of this instantaneous phase error and the fractional spurs can be achieved, however. This is because the contents of the accumulator approximate the instantaneous phase error and can, if properly scaled and subtracted from the phase detector output, compensate for the phase jitter. Such cancellation is done by internal circuitry in the UMA1005 synthesizer.

As different channels are selected the fractional word is updated and the main fractional spur moves relative to the carrier. For the higher fractional words the spur is at such wide offsets (150, 180, 210 kHz in our example) that the closed loop response effectively eliminates it. The closest the main spur can get is within one frequency step of the carrier (30 kHz in our example) so ensuring good spurious performance means the same characterization as with a standard synthesizer.

The use of an adaptive loop, having a narrower loop bandwidth for a given switching speed than a nonadaptive loop, will ensure that maximum fractional and comparison frequency spurious rejection is obtained.

We now return to the single-synthesizer design challenge presented earlier, armed with this new synthesizer. In order to see the benefits of fractional-N synthesis a simple non-adaptive design was constructed that closely matched the MB1501 design above. The same VCO, channel spacing, TCXO quality and closed loop bandwidth were chosen. Only the synthesizer and loop filter values were changed.

Figures 6 and 7 show the loop filter and resulting close-in noise spectrum. The SSB phase noise at a 1 kHz offset was:

$$
\begin{aligned}
L(1\text{ kHz}) &= -79.5\text{ dBm} - (-5\text{ dBm}) \\
&= -74.5\text{ dBc/Hz}
\end{aligned}
$$

This easily met the target specification, as did the 30 kHz spurious of –62 dBc.

Study of the close-in noise response in comparison with the MB1501 shows some 15.8 dB improvement in the SSB phase noise due to the lower division ratio, as would be expected. It also should be noted that, although the two designs show roughly the same closed loop bandwidth the UMA1005 shows less peaking. This is to be expected as the phase comparator is operating at 240 kHz and this is much higher than the closed loop bandwidth. The additional phase margin and increased comparison frequency reduced the switching time (Figure 8) to 1.0 ms, some 33 percent faster than the MB1501 design.

Finally, we note that the fractional compensation is working quite well, as the fractional spurious with the lowest fractional word (30 kHz offset, NF=1) was 10 dB better than the MB1501 design (Table 2).

## Adaptive Fractional-N Design

As mentioned above, it is possible to improve the noise and spurious response, and decrease the switching time using an adaptive design. Such a design was made using the procedure in the UMA1005 application note. This was designed for a closed loop bandwidth of 1.5 kHz in normal mode and 4.5 kHz in

| | | MB1501 | UMA1005 | |
|---|---|---|---|---|
| VCO frequency | $f_{VCO}$ | 918.03 | 918.03 | MHz |
| Channel spacing | $f_{SP}$ | 30 | 30 | kHz |
| Comparison frequency | fC | 30 | 240 | kHz |
| Overall Division Ratio | N | 30,601 | 3825.125 | |
| SSB Phase Noise | L(1 kHz) | –58.7 | –74.5 | dBc/Hz |
| 30 kHz sidebands | \|30 kHz\| | –52 | –62 | dBc |
| 240 kHz sidebands | \|240 kHz\| | — | –76 | dBc |
| Switching time | $t_S$ | 1.5 | 1.0 ms (w/in 1 kHz) | |

**Table 2. Comparison of UMA1005 with MB1501.**

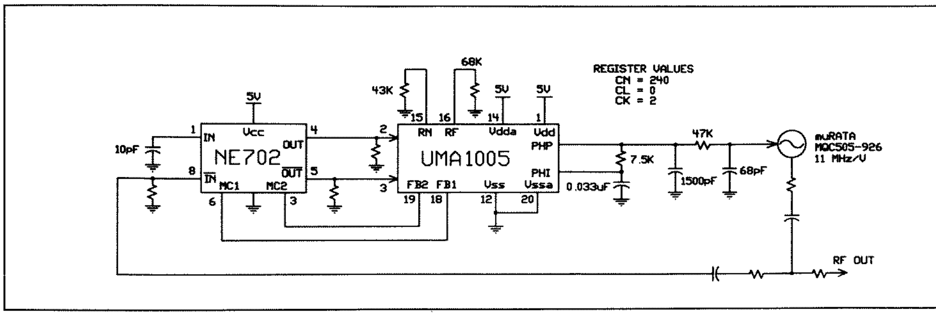



**Figure 9. Adaptive UMA1005 design schematic and register values.**

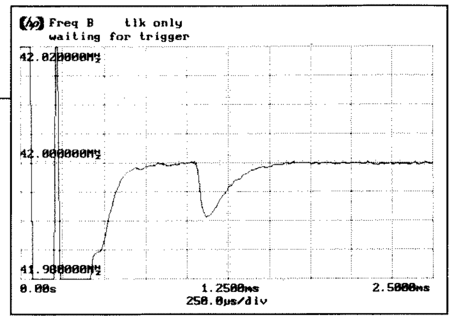



**Figure 11. Switching response of adaptive UMA1005 design (5 kHz/division, output signal mixed down to 42 MHz).**

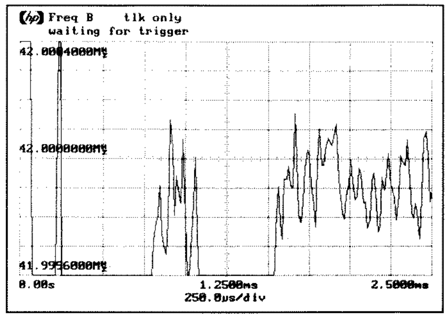



**Figure 12. Switching response of adaptive UMA1005 design (100 Hz/division, output signal mixed down to 42 MHz).**

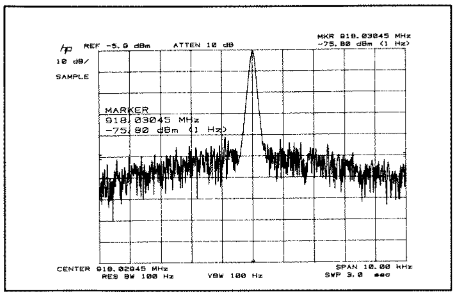



**Figure 10. Close-in noise spectrum of adaptive UMA1005 design.**

speed-up mode. The resulting schematic and register values are given in Figure 9. The close-to-carrier output spectrum is shown in Figure 10, which reveals the closed loop bandwidth of about 1.5 kHz and SSB phase noise of:

$$L(1\text{ kHz}) = -75.8\text{ dBm} - (-6.9\text{ dBm}) = -68.9\text{ dBc/Hz}$$

This yielded residual FM of 77 Hz RMS, unweighted, from 50 to 15000 Hz, and integrated jitter of 1.6 degrees from 200 to 20000 Hz.

The fractional sidebands were reduced to –68.5 dBc at 30 kHz, some 13.5 dB better than the requirement. For the next channel (NF = 2) the 60 kHz fractional spur was at –86 dBc. The comparison frequency feedthrough at 240 kHz was always below –85 dBc.

Figures 11 and 12 show the switching response to a 15 MHz output frequency step. The design meets both the coarse (1 kHz) and the fine (100 Hz) switching time requirements. It should be noted that the fine settling could be improved somewhat as adaptation time was arbitrarily set at 1 ms, at which point changing loop gain causes a predictable glitch of a few kilohertz. Reducing the time in the wider bandwidth mode should reduce the final settling time by about 0.3 ms.

## Basic Design Procedure

The following procedure will yield a first-pass loop filter design for the UMA1005/1014/1016 PLL circuits (including nonadaptive designs using the UMA1005 - adaptive designs are documented in the application note for that device):

a. Select basic loop and synthesizer parameters:

- output signal frequency, $f_{VCO}$, Hz
- VCO gain, $K_V$, Hz/V
- switching time, $t_S$, seconds
- comparison frequency, $f_C$, Hz (equals channel spacing using UMA1014/1016, 5 or 8 times channel spacing using UMA1005)
- loop damping factor, zeta, unitless (0.707 = fastest switching, 1.0 = low overshoot)
- charge pump output current, $I_{CP}$, Amperes (selectable using UMA1005/1014; fixed value using UMA1016)

b. Calculate approximate natural frequency of loop from switching time requirement:

- natural frequency, $f_N \approx 2/t_S$

c. Calculate overall division ratio and the most critical component values:

- loop division ratio, $N = f_{VCO}/f_C$
- main capacitor $C_3 = K_V\, I_C/[N(2\pi f_N)^2]$
- damping resistor $R_2 \approx 2\,\text{zeta}\,[N\,(K_V\, I_{CP}\, C_3)]^{0.5}$
- filter capacitor $C_1 = C_3/10$

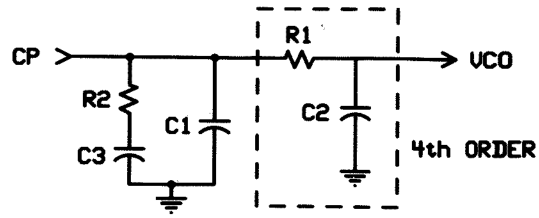
## PLL Simulation and Optimization

The above component values will yield a third order system response from a second order filter response (with one pole due to the VCO). The resulting values are only approximations, however, as they were derived by assuming the loop was a second order type. For more accurate analysis and optimization, key loop parameters and the filter response can then be entered into a software analysis program to verify loop stability and switching time, such as (4). This product is for use on MS-DOS computers and includes graphical outputs of open loop, closed loop, switching time and error responses, as well as having the ability to simulate phase noise. The formulas in Figure 13 can be used to convert component values into parameters requested by that program.

*Warning* — Although the gain of phase detectors is most commonly specified in volts/radian, circuits (such as the UMA1005/1014/1016) that have integrated charge pumps are typically specified in output current (amperes or amperes/cycle, used interchangeably). Remember to use $I_{CP}$ in amperes for calculation of loop filter components and use the formula in Figure 13 to convert the $I_{CP}$ value and loop filter component values to a single Volts/radian value for purposes of the PLL analysis program.

Study of the Figure 13 formulas shows that $C_3$ is the single most critical filter component determining the loop dynamics, as it determines not only the dominant zero frequency but also the detector gain. Be sure to perform the analysis using the standard capacitor value to be built into the circuit. As this is the largest capacitor in the circuit, attention to the type of dielectric used will help minimize switching time and leakage (see Tips from the Trenches below). As the UMA1005 has programmable charge pump currents, adjustable with 8-bit resolution, the gain is easily adjusted to optimize loop performance once a particular $C_3$ value is chosen.

This linear analysis does not include calculation of comparison frequency sideband magnitudes arising from nonlinear effects. After a design has been analyzed to be stable, it should be prototyped and the sidebands measured. If they are not suppressed sufficiently then the additional lag network ($R_1$, $C_2$) can be added. This network will change the




**Loop Parameters**

| Parameter | 3rd Order | 4th Order | Units |
|---|---|---|---|
| Kd | $\frac{I_{cp}(A/cycle)}{2\pi(C_1+C_3)}$ | $\frac{I_{cp}}{2\pi(C_1+C_2+C_3)}$ | V/rad |
| Poles (1st Order) | 2 at 0<br>1 at $\frac{C_1+C_3}{2\pi R_2C_1C_3}$ | 2 at 0 | Hz |
| Pole (2nd Order) | N/A | $\frac{1}{2\pi}\sqrt{\frac{C_1+C_2+C_3}{R_1R_2C_1C_2C_3}}$ | Hz |
| Damping | N/A | $\pi f_p\left[\frac{R_2C_3(C_1+C_2)+R_1C_2(C_1+C_3)}{C_1+C_2+C_3}\right]$ | - |
| Zero (1st Order) | $\frac{1}{2\pi R_2C_3}$ | $\frac{1}{2\pi R_2C_3}$ | Hz |

**Figure 13. Basic passive loop filter design.**

system to a fourth order type, with the resulting frequencies and damping factor listed in Figure 13. It is important to keep the second order pole frequency much higher (typically at least a decade above) the dominant zero frequency in order to ensure adequate phase margin. Be sure to include the VCO internal capacitance (add to $C_2$ value) when calculating the second order pole frequency.

Before prototyping, the simulated magnitude of the closed loop response at the comparison frequency (with the extra pole added) can be compared to the original circuit response to verify the additional amount of sideband suppression to be expected. The gain and phase margins will be reduced somewhat. If placed at too low a frequency both the chances of oscillation and the final settling time will increase.

## Tips From the Trenches

Many years ago a wise applications engineering manager advised me that "Q doesn't come in small packages". This axiom is very pertinent to synthesizers and affects the selection of several of the external components. The words ring in my ears with each ring of the telephone from someone attempting to ignore this advice.

Much has been written on the importance of VCO and TCXO oscillator quality but I will emphasize it here for the synthesizer novice: Use low-grade oscillators and your phase noise (and more) will suffer. Smaller usually means lower quality. The TCXO phase noise dominates the phase noise closest to the carrier, typically at offsets below 100 Hz. The VCO phase noise is essentially the system phase noise at offsets of greater than a few times the loop bandwidth and therefore can affect integrated jitter, adjacent channel protection and intermodulation performance.

Probably one of the less discussed components is the large integrator capacitor in the loop filter. Due to the large values often required, there is considerable temptation to keep size small by using Z5U or X7R ceramic or, worse, electrolytic types. The problems with these include higher leakage current, which raises spurious levels, and more dielectric absorption. This second factor is "memory" in the capacitor dielectric that can really degrade the settling time. The switching time data in all of the above designs were taken using polyester film type capacitors. Figure 14, when compared to Figure 8, shows how the use of an X7R ceramic capacitor will

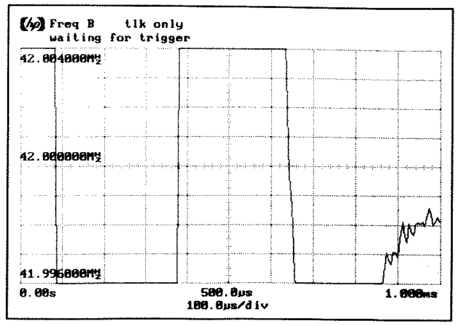



**Figure 14. Effect of X7R capacitor on switching response.**

increase final settling time. Polypropylene and polystyrene are the best but, for non-military budgets, polyester film types provide much improvement at a reasonable cost.

## Summary

The UMA1005 has features making it especially attractive for emerging digital communications markets, where simultaneous requirements for fast hopping and low noise pose challenges not met by typical low cost commercial devices. The UMA1005 is in production now, with prices from $5.81/500s. For more information, call (800) 447-1500, ext. 1008, or circle Info/Card #161. *RF*

## References

1. U. Rohde, "Low-Noise Frequency Synthesizers Using Fractional-N Phase-Locked Loops," *RF Design*, Jan./Feb. 1981.
2. U. Rohde, *Digital PLL Frequency Synthesizers*, Prentice-Hall, 1983.
3. R. Hassun, "Common-Denominators in Fractional-N," *Microwaves & RF*, June 1984.
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## About the Author

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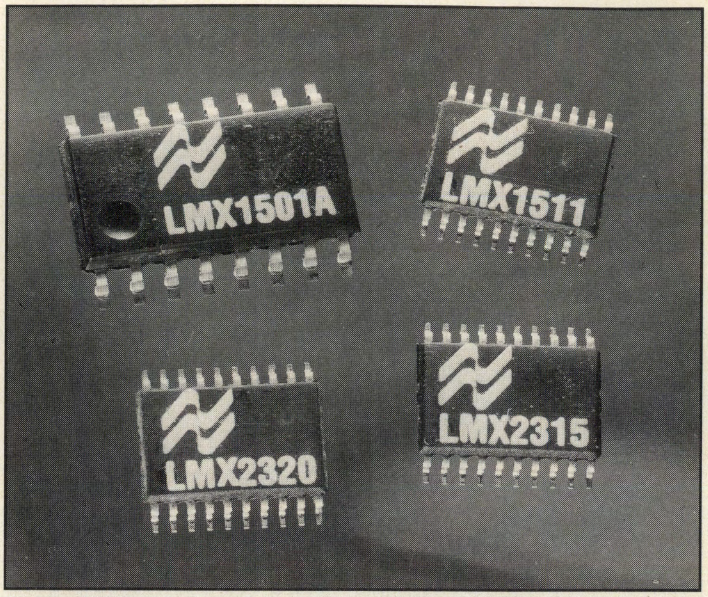
# Radio Transceiver ICs

National Semiconductor has announced the PLLatinum™ series of phase locked loop (PLL) radio frequency synthesizers. The PLLatinum series is the first in a line of silicon radio frequency (SiRF™) solutions to be introduced by National Semiconductor for the radio transceiver marketplace. Initial members of the PLLatinum series are the LMX1501A, LMX1511, LMX2315 and LMX2320. The LMX-1501A/1511/2315 all operate to 1.1 GHz with low power consumption, 18 mW at 3 V. All can operate from 2.7 to 5.5 Volts, have fast lock times, low phase noise, a balanced charge pump and dual modulus scaler (64/65 or 128/129). The LMX2320 operates to 2.0 GHz with 36 mW power consumption. Power consumption of the LMX2320 drops to less than 3 mW when in the power down mode. The LMX-2315 shares this power down feature and, along with the 2320 and 1511, is packaged in the super thin (1.1 mm) TSSOP package. A dual bandwidth switching function for frequency scanning applications is part of the LMX1511, 2315 and 2320. The LMX1501A is available in JEDEC and EIAJ SO packages and is pin compatible with Fujitsu's MB1501. The PLLatinum series is in pre-production and will begin high-volume production in March 1993. Pricing starts at $4.25 per unit, in quantities of 1000.
**National Semiconductor Corp.**
**INFO/CARD #250**



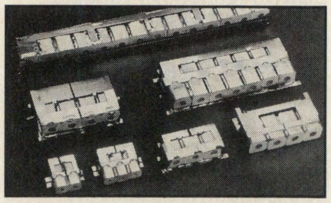

## Synthesized Sweep Generator

Wiltron has announced the Wiltron 68100A Series synthesized sweep generators. This series offers high performance and economy for microwave stim-

ulus requirements. Four broadband models meet the needs of network analysis and L.O. replacement applications — the 68137A (2 to 20 GHz), the 68147A (0.01 to 20 GHz), the 68163A (2 to 40 GHz) and the 68169A (0.01 to 40 GHz). All of these frequency synthesized models have single sideband noise less than –88 dBc/Hz at 10 kHz offset from 10 GHz and spurious responses typically less than –60 dBc. All models possess complete analog/digital/manual sweep capability, and AM, FM and square wave modulation via external modulating signals. Output power is +13 dBm to 20 GHz and +6 dBm to 40 GHz. The 68100A series has an easy-to-use menu-driven interface and remote operation capability via an IEEE-488 interface. The front panel interface can be omitted for dedicated and A.T.E. applications. Prices for the 68100A Series start at $23,000.
**Wiltron Co.**
**INFO/CARD #249**

## Custom Ceramic Filter Manufacturing

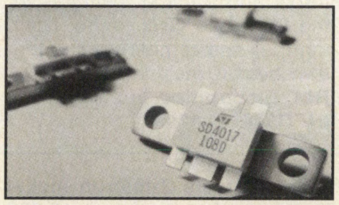
Integrated Microwave offers low cost, custom, high Q ceramic filters for a wide variety of applications, including GPS, cellular (U.S. and international), PCN, LAN, CATV, INMARSAT and ISM band applications. Combline, coaxial and other filter types are available. These custom filters are temperature stable (5 ppm/degree C), low loss designs available with Chebyshev, elliptic, linear phase and Butterworth transfer functions. Two to eight-section filters are available, and frequencies range from 300 to 6000 MHz. 3 dB bandwidths



range from 1 to 10 percent in coaxial type filters and 1.0 to 20 percent in combline types. Units can be manufactured in surface mount, PC mount or connectorized configurations with a variety of package designs. Hermetically sealed units are also available. Quantities can range from single prototypes to production quantities, and delivery time is typically four weeks or less.
**Integrated Microwave**
**INFO/CARD #248**

## RF Power Transistor

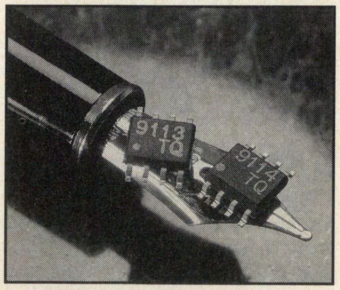
SGS-Thomson Microelectronics has introduced a new silicon bipolar transistor that is particularly suitable for use in Class AB linear power amplifiers for digital



cellular telephone base stations. Known as the SD4017, the device is a common emitter, input-matched power device that provides 30 W minimum output power with 7.5 dB minimum power gain. Collector efficiency is typically 55 percent, DC supply voltage is 25 V and the device is supplied in an industry standard 0.230-inch, 6-lead flanged package. The integral input impedance matching network optimizes the wideband capability, allowing the operating range to be located in any part of the 806-960 MHz band. In addition to the highly linear characteristics required for cellular telephone applications, the SD4017 features refractory gold metallization and diffused emitter ballast resistors, ensuring low thermal resistance and long-term reliability under Class AB operation.
**SGS-Thomson Microelectronics**
**INFO/CARD #247**

## IF/AGC MMIC Amplifiers

TriQuint Semiconductor has released two additions to its family of MMIC downcoverter building blocks. The TQ9113N and TQ9114N IF/AGC amplifiers operate between 30 and 500 MHz and are packaged in the industry standard, low-cost SO-8 package. Current consumption is a low 2.2 mA/3.1 mA (TQ9113N/TQ9114N) from a single +5 V supply. Up to 30 dB of gain is available with a wide AGC range of 65 dB (TQ9114N), which

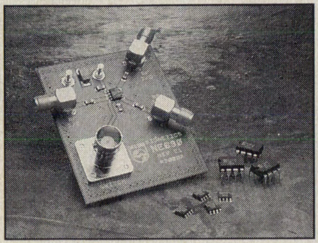


simplifies circuit design compared to a discrete approach. Noise figure ranges from 4.5 to 6.0 dB. Products are available from stock in quantities less than 1000 units, with orders greater than 1000 quoted at four weeks ARO. Pricing in quantities of 500 to 999 is $5.25 for the TQ9113N and $5.60 for the TQ9114N. These products are available in shipping tubes or tape and reel. Detailed data sheets are available.
**TriQuint Semiconductor, Inc.**
**INFO/CARD #246**

# Product Spotlight: Switches

## Low Power Switch

The NE/SA630 from Philips Semiconductors is the industry's lowest power consumption radio-frequency switch. The device is a bi-directional, single-pole, double-throw switch that



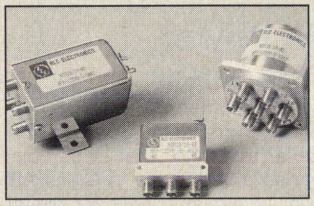
handles wideband signals ranging from DC to 1 GHz. The NE/SA630 is packaged in 8-pin DIP and SOIC packages and draws just 140 uA from a 5 V supply. No external components are necessary and typical switching times are 25 ns.
**Philips Semiconductors**
**INFO/CARD #245**

## Switch Matrices

The PX series of switch matrices from Cytec requires only 3.5 inches of rack space and has 512 single or two-pole relays pre-wired in the required matrix configuration. Individual chassis can be interconnected to form matrices of any size. High bandpass goes to 30 MHz and crosstalk and isolation are better than –60 dB at 1 MHz. The PX/512 Mainframe is priced at $1000, the IF-4 IEEE- 488 control module is $500, and individual 2x16 switch modules are $300 apiece.
**Cytec Corp.**
**INFO/CARD #244**

## QPL Listed Coaxial Switches

RLC Electronics' line of coaxial switches contains over 75 models which are qualified to MIL-S-3928 Revision D. These switches included SPDT, multi-



throw and transfer switches in the DC to 20 GHz range. Both manual and remotely actuated switches are included.
**RLC Electronics, Inc.**
**INFO/CARD #243**

## Multi-Throw Switch for Test Equipment

Dow-Key Microwave has developed a series of DC to 18 GHz multi-throw switches designed for use in compact switch matrices. The 465-Series switch is compatible with either 12 or 28 V drivers, and meets MIL-STD 3928. Among the available options are: radial or in-line designs, single-pole to 12 throw, indicator contacts and soldered or D type control terminals.
**Dow-Key Microwave Corp.**
**INFO/CARD #242**

## High Power Transfer Switch

The model FS50 Series from Sage Laboratories is designed for cellular base station applications, with excellent performance specifications from DC to 12. 4 GHz. The fail-safe transfer switch is nomally closed. The switch handles 500 W CW at 1 GHz and employs a break-before-make switching mechanism, to prevent hot switching.
**Sage Laboratories, Inc.**
**INFO/CARD #241**

# TOOLS, MATERIALS & MANUFACTURING

## Antenna Members

Braiding technology and resin transfer molding are being used to manufacture lightweight, corrosion-resistant structural members for antenna systems. Braided composite members are transparent to radio waves and are non-magnetic. They are 40 percent lighter than steel members with the same strength ratings but do not interfere with reception or transmission.
**Bentley Harris**
**INFO/CARD #240**

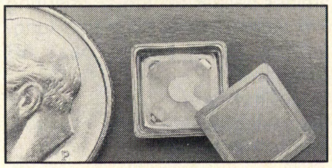
## Conductive Filler

High performance with lower loading concentration can be obtained with very fine nickel fiber. Ribtec's 2-micron nickel fiber has a length of over 10,000 miles per pound. This allows a lower concentration of nickel to form a highly conducting network with shielding effectiveness that can exceed 90 dB in the 30 to 1000 MHz range.
**Ribbon Technology Corp.**
**INFO/CARD #239**

# DISCRETE COMPONENTS

## SMT Crystals

EG&G Frequency Products announces a new AT cut crystal in a surface mount package with standard motional parameters available. This package contains



the same high performance features found in EG&G's complete line of AT-cut crystals, only smaller. Theses surface mount crystals meet MIL-STD-202 method 210, resistance to soldering heat.
**EG&G Frequency Products**
**INFO/CARD #238**

## Power Resistors

Caddock model MP808 and Caddock model MP816, are Kool-Pak™ non-inductive power film resistors in TO-126 and TO-220 cases, respectively. The MP808 features 8 Watt dissipation at case temperatures of 25 degrees C, while the MP816 dissipates 16 Watts at 25 degrees C. Both employ a thermally conductive, molded package.
**Caddock Electronics, Inc.**
**INFO/CARD #237**

## Thin SMT Crystals

By combining a smaller crystal blank, a new header base and a new cold-sealing technique, Raltron Electronics has been able to shrink the thickness of the model H-13 to 1.3 mm. The H-13's frequency stability is rated at ±50 ppm at 25 degrees C, and shunt capacitance is 7.0 pF max. The crystals are available from 16 to 100 MHz. Depending on frequency, prices range from $3.00 to $4.00 each in quantities of 1000.
**Raltron Electronics Corp.**
**INFO/CARD #236**

## Resistors and Capacitors

A new range of X7R and NPO ceramic multilayer chip capacitors and a new line of 1 Watt chip resistors have been released by Philips Components. The new capacitors are rated at 200 V and are available in four sizes: 0805, 1206, 1210 and 1812. The PRC201 SMT power resistors come in 1218 packaging, which has the same dimensions as 1812 devices, but with terminations on the longer sides.
**Philips Components**
**INFO/CARD #235**

# SEMI-CONDUCTORS

## Voltage-Feedback Op Amp

Burr-Brown's OPA622 is a high-speed amplifier which can be used in either current or voltage-feedback modes. Specifications include 200 MHz large-signal bandwidth at 5 Vp-p output swing, 1700 V/us slew rate and 78 dB common-mode rejection. The OPA622 operates from a ±5 V supply and is priced from $7.10 in 100s.
**Burr-Brown Corporation**
**INFO/CARD #234**

## Clock Recovery Chip

Designed for fiber optic and data links at standard DS-3 and OC-1/STS-1 rates, Analog Devices' AD800-45 and AD800-52 provide clock recovery and retiming functions in compact 20-pin devices that require just a single external capacitor for operation. The AD800-45 (44.736 Mbps) and the AD800-52 (51.84 Mbps) are $30 in the 1000s.
**Analog Devices, Inc.**
**INFO/CARD #233**

## Broadband Amplifier

The model AC3055 thin-film cascadable amplifier operates over the frequency range of 0.01 to 3 GHz. Typical gain is 10.5 dB. Maximum noise figure is 3.0 dB from 1.5 to 3.0 GHz, 3.5 dB from 0.75 to 1.5 GHz and 4.5 dB from 0.2 to 0.75 GHz over the temperature range of 0 to 50 degrees C. Operating current is 56.0 mA at 5 V.
**Cougar Components**
**INFO/CARD #232**

## 600 W, Gold Transistor

Motorola has introduced the MRF157 RF power transistor, the industry's first all gold 600 Watt device of its kind. Operating with a supply voltage of 50 V, typical performance characteristics include gain of over 20 dB at 30 MHz, drain efficiency of 45 percent and a 3rd order IMD at rated output power of –25 dB typ. Pricing for low volume quantities of the MRF157 is $499.
**Motorola, Inc.**
**INFO/CARD #231**

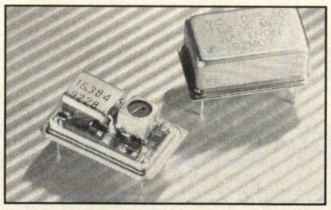
## GaAs FET Voltage Regulator

The R-15105 is a negative and positive regulator ideally suited for biasing GaAs FET amplifiers or any system or sub-assembly that requires negative and positive voltages. Standard outputs are –5.0 VDC at 100 mA and +10.0 VDC at 6.0 A with input voltage of ±15 VDC. The regulator is thermal and current overload protected and will shut off positive output voltage when negative input voltage is removed.
**RF Power Components, Inc.**
**INFO/CARD #230**

# SIGNAL SOURCES

## Fast VCXOs

Raltron Electronics has announced a family of high performance, voltage-controlled crystal oscillators. The VC-7000 series has a frequency range of 500 kHz to 150 MHz, high deviation sensitivity and squarewave output with rise/fall times less than 10 ns. Output drive is specified at 10 TTL loads or 15 pF for



HCMOS loads. Units operate from +5 V and draw 35 mA. The VC-7000 series VCXOs are priced at $15.00 each in the 1000s.
**Raltron Electronics Corp.**
**INFO/CARD #229**

## Digitally Tuned Oscillator

Model 2232 digitally tuned oscillator (DTO) has a frequency range of 2 to 4 GHz. The oscillator is PROM linearized and stabilized by an internal heater to provide a high degree of tuning repeatability. Minimum output power is 10 dBm with less than 4 dB p-p power variation. Settling time is 1 us with a tuning resolution (12 bit input) of 0.5 MHz. Other models up to 20 GHz are available.
**Radian Technology, Inc.**
**INFO/CARD #228**

## Compact VTCXO

Philips Components has launched a voltage-controlled, temperature compensated crystal oscillator occupying only 0.6 cm$^3$. Covering 8 to 20 MHz, the device meets reference oscillator requirements for both U.S. ADC (19.44 MHz) and European GSM (13 MHz) cellular systems. The device comes in a 4-pin sealed metal can which is pin compatible with DIL-8 corner-pinning.
**Philips Components**
**INFO/CARD #227**

# SUBSYSTEMS

## Data Transceiver

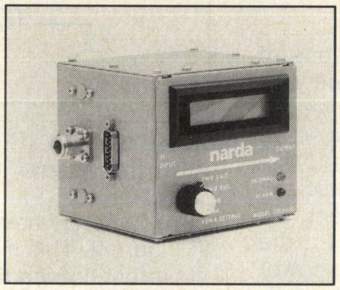
The KED TR-1 is a complete two way radio used for short distance wireless transmission of digital data. Distances achieved indoors are in excess of 50 feet at data rates of 2.4 kbaud. The transceiver works in half duplex mode. It is low power and very small. It is FCC part 15 certifiable as part of a product and is easy to incorporate into designs.
**Kiefer Electronic Development**
**INFO/CARD #226**

## Arabsat LNA/ Downconverter

The Arabsat-Pro LNB is a low noise amplifier/downconverter which operates on the 2.50 GHz to 2.70 GHz input frequency. The unit is used to receive S-band transmissions in the areas served by the Arabsat hybrid (C-band & S-band) satellites. The product, part no. 31692, features an exceptionally low noise figure of 0.60 dB (typ.) and 62 dB gain (typ.).
**California Amplifier**
**INFO/CARD #225**

## Power/VSWR Monitor

Narda's CellGuard was designed for long-term system monitoring of both combined transmitter power and antenna




feed circuit (VSWR) performance. Model CEL8450 measures true RMS transmitted power for one, two or up to 16 transmitters using the same antenna. An RS-485 interface allows remote monitoring, alarm threshold input and alarm resetting.
**Loral Microwave-Narda**
**INFO/CARD #224**

# SIGNAL PROCESSING COMPONENTS

## Linearized Variable Attenuator

Amplifonix has announced the model TGLR9025 voltage variable attenuator with an integral linearizer. The linearity is typically better than ±2.0 dB over the attenuation range of 6.0 to 48 dB. The attenuation slope is 5 dB per volt over the control voltage of –1 to –10 VDC. The unit is available in TO-8B, flatpack and 0.625 inch square gull-wing SMT packages.
**Amplifonix, Inc.**
**INFO/CARD #223**

## Dual Directional Coupler

Model DC-312 is a dual directional coupler covering the frequency range of 225 to 400 MHz with a minimum of 18 dB directivity. Nominal coupling values are 11 and 20 dB, with loss of 1.2 and 0.5 dB, respectively. The coupler measures 2.10 x 1.75 x 0.40 inches with snap microstrip line launch transition.
**Microwave Communications Laboratories, Inc.**
**INFO/CARD #222**

## Hermetic Filters

AVX Filters' hermetic solder-in filter line includes the smallest hermetic filter of its type. These filters meet MIL-F-28861 specifications. These low-pass filters for medium to high frequency applications can withstand 400 degree C installation temperatures. The filters are available in standard industry sizes plus custom and multiple filter arrays.
**AVX Filters**
**INFO/CARD #221**

## High Power Coupler

Narda's dual 50 dB dual directional coupler provides excellent directivity and power handling. Model CEL30243 offers forward and reflected –50 dB coupled power ports and a minimum 25 dB directivity, with 30 dB or more directivity typical. The CEL30243 is conservatively rated at 600 W CW input power.
**Loral Microwave-Narda**
**INFO/CARD #220**

## Limiting Filters

GEC-Marconi Research is now offering a range of frequency selective limiting filters. They feature pass bandwidths up to 1 GHz within the range of 1.5 GHz to 4 GHz and typically exhibit a frequency selectivity of better than 10 MHz for up to 30 dB compression of high level signals. Limiting power thresholds between –10 dBm an –20 dBm are available with nominal 3 dB variation across the passband.
**GEC-Marconi Research**
**INFO/CARD #219**

# A Clearer Derivation of the Microwave Gain Equations

By Dean A. Frickey

*In microwave amplifier design and analysis, a clear understanding of S-parameters and the gain equations is required. The derivation of these gain equations is often gone over lightly in text books. In addition, most books will show,*

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0} \qquad \Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

*yet fail to state that these equations require $Z_0$ to be real.*

This paper begins with definitions of incident and reflected current and voltage and the definitions for the a's and b's used in the S-parameter equations. From these are derived expressions for the input reflection coefficient, $\Gamma_1$, the output reflection coefficient, $\Gamma_2$, the source reflection coefficient, $\Gamma_S$, the load reflection coefficient, $\Gamma_L$, the power relations in terms of a's and b's from S-parameters, and the gain relationships. In doing this, no assumptions have been made requiring real normalizing impedances so the results are complete.

### Incident and Reflected Components

The terms incident and reflected, while they conjure up pictures in our minds, need to be defined in terms of circuit theory. Following the work of Carson (1), the circuit of Figure 1 provides the definitions for the incident voltage and current. The incident current, $I_i$, is defined as the current through the load when it is conjugately matched to the source. From the circuit in Figure 1, this gives,

$$I_i = \frac{V_S}{Z_0 + Z_0^*} \tag{1}$$

The incident voltage, $V_i$, is defined as the voltage across the load when it is conjugately matched to the source. It is given by,

$$V_i = \frac{V_S Z_0^*}{Z_0 + Z_0^*} \tag{2}$$

where * indicates complex conjugate.

When the actual load is connected to

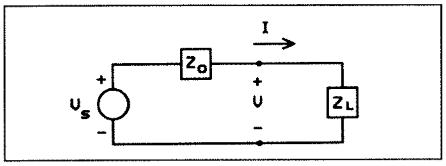



**Figure 1. Circuit for the definition of the incident current and voltage.**

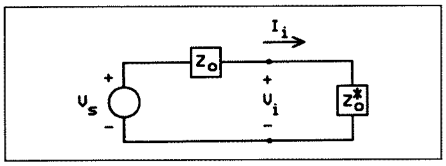



**Figure 2. Voltages and currents with the actual load.**

the source we have the situation shown in Figure 2. This circuit defines the actual current flowing through and the actual voltage across the load. Here, the actual current is given by,

$$I = \frac{V_S}{Z_0 + Z_L} \tag{3}$$

The actual voltage across the load is,

$$V = \frac{V_S Z_L}{Z_0 + Z_L} \tag{4}$$

The reflected current, $I_r$, is defined as the difference between the incident current and the current to the actual load.

$$I_r = I_i - I \tag{5}$$

Then,

$$I = I_i - I_r \tag{6}$$

In a similar manner the reflected voltage, $V_r$, is defined as the difference between the voltage across the load and the incident voltage.

$$V_r = V - V_i \tag{7}$$

Then,

$$V = V_i + V_r \tag{8}$$

Since $Z_0$ is used in the definitions of the voltages and currents, it is referred to as the normalizing impedance. $Z_0$ will turn up in the S-parameter derivations and it is important to know where it comes from.

We need to get relationships between the incident voltage and current, and between the reflected voltage and current. To do this, substitute equations 1 and 3 into equation 5 to get,

$$I_r = \frac{V_S (Z_L - Z_0^*)}{(Z_0 + Z_0^*)(Z_0 + Z_L)} \tag{9}$$

Substituting equations 2 and 4 into equation 7 gives

$$V_r = \frac{V_S Z_0 (Z_L - Z_0^*)}{(Z_0 + Z_0^*)(Z_0 + Z_L)} \tag{10}$$

These expressions, along with equations 1, 2, 3, and 4 give the voltage and current components in terms of the source voltage, the normalizing impedance, and the load impedance.

From equations 1 and 2, we have

$$V_i = Z_0^* I_i \tag{11}$$

and from equations 9 and 10, we have

$$V_r = Z_0 I_r \tag{12}$$

Equations 11 and 12 will be used in the derivations to follow.

The equations in this section were developed for a 1-port network. By simply adding subscripts to the voltages and currents, the equations are valid for multi-port networks. In this paper, expressions are being developed for 2-port networks so the remaining derivations will only refer to 2-ports.

### Scattering Parameters

Now, for a 2-port network, the scattering parameters are defined by the relationships,

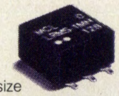
$$b_1 = S_{11} a_1 + S_{12} a_2 \tag{13a}$$

$$b_2 = S_{21} a_1 + S_{22} a_2 \tag{13b}$$

a and b are defined as,

$$a_j = \left[\frac{Z_{oj} + Z_{oj}^*}{2}\right]^{\frac{1}{2}} \cdot I_{ji} \tag{14}$$

$$b_j = \left[\frac{Z_{oj} + Z_{oj}^*}{2}\right]^{\frac{1}{2}} \cdot I_{jr} \tag{15}$$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| Model | Freq. Range (MHz) LO. RF | IF | LO Level (dBm) | Conv. Loss Mean ($\bar{X}$) mid-band (dB) | L-R Isol. Mean ($\bar{X}$) mid-band (dB) | Price $ (1-9) |
|---|---|---|---|---|---|---|
| LRMS-2L | 800-1000 | DC-200 | +3 | 6.6 | 24 | 6.95 |
| LRMS-1 | 0.5-500 | DC-500 | +7 | 6.4 | 45 | 6.25 |
| LRMS-1W | 2.0-750 | DC-750 | +7 | 5.8 | 45 | 6.75 |
| LRMS-2 | 5-1000 | DC-1000 | +7 | 6.8 | 38 | 6.95 |
| LRMS-2D | 5-1000 | DC-1000 | +7 | 6.8 | 40 | 7.25 |
| LRMS-2U | 10-1000 | 10-1000 | +7 | 6.5 | 46 | 11.45 |
| LRMS-5 | 5-1500 | DC-1000 | +7 | 6.0 | 41 | 13.95 |
| LRMS-11A | 1400-1900 | 40-500 | +7 | 7.0 | 25 | 16.95 |
| LRMS-1LH | 2.0-500 | DC-500 | +10 | 5.8 | 47 | 7.95 |
| LRMS-2LH | 5-1000 | DC-1000 | +10 | 6.6 | 40 | 8.95 |
| LRMS-5LH | 10-1500 | DC-900 | +10 | 5.4 | 38 | 14.95 |
| LRMS-1MH | 2.0-500 | DC-500 | +13 | 5.7 | 44 | 8.95 |
| LRMS-2MH | 5-1000 | DC-1000 | +13 | 6.6 | 44 | 9.95 |
| LRMS-5MH | 10-1500 | DC-900 | +13 | 5.8 | 46 | 15.95 |
| LRMS-1H | 2.0-500 | DC-500 | +17 | 6.3 | 44 | 10.95 |
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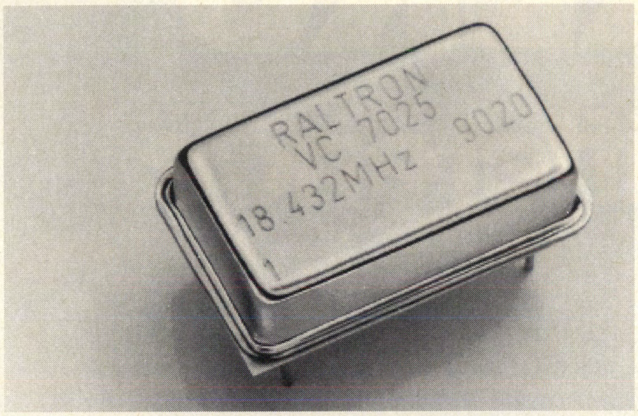


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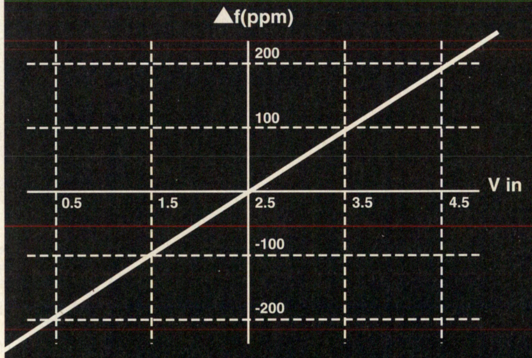


**Crystals**
**Crystal Oscillators**
**Crystal Filters**
**Ceramic Resonators**

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where $Z_{0j}$ is the normalizing impedance for the jth port.

We want to find expressions for a and b in terms of the port voltages and currents and normalizing impedances. To begin, solve equations 14 and 15 for $I_i$ and $I_r$.

$$I_{ji} = \left[\frac{2}{Z_{0j} + Z_{0j}^*}\right]^{\frac{1}{2}} \cdot a_j \qquad (16)$$

$$I_{jr} = \left[\frac{2}{Z_{0j} + Z_{0j}^*}\right]^{\frac{1}{2}} \cdot b_j \qquad (17)$$

Then, substituting them into equation 6 gives,

$$I_j = \left[\frac{2}{Z_{0j} + Z_{0j}^*}\right]^{\frac{1}{2}} (a_j - b_j) \qquad (18)$$

Using equations 11 and 12 in equation 8 gives,

$$V_j = Z_{0j}^* I_{ji} + Z_{0j} I_{jr} \qquad (19)$$

Then substituting equations 16 and 17 into equation 19 gives,

$$V_j = \left[\frac{2}{Z_{0j} + Z_{0j}^*}\right]^{\frac{1}{2}} (a_j Z_{0j}^* + b_j Z_{0j}) \qquad (20)$$

Multiply equation 18 by $Z_{0j}$ and add it to equation 20. This gives,

$$V_j + Z_{0j} I_j = (Z_{0j} + Z_{0j}^*)\left[\frac{2}{Z_{0j} + Z_{0j}^*}\right]^{\frac{1}{2}} \cdot a_j$$

Solving for $a_j$ gives,

$$a_j = \frac{V_j + Z_{0j} I_j}{\left[2(Z_{0j} + Z_{0j}^*)\right]^{\frac{1}{2}}} \qquad (21)$$

If now we multiply equation 18 by $-Z_{0j}^*$ and then add to equation 20 we get,

$$V_j - Z_{0j}^* I_j = (Z_{0j} + Z_{0j}^*)\left[\frac{2}{Z_{0j} + Z_{0j}^*}\right]^{\frac{1}{2}} \cdot b_j$$

Solving for $b_j$ gives,

$$b_j = \frac{V_j - Z_{0j}^* I_j}{\left[2(Z_{0j} + Z_{0j}^*)\right]^{\frac{1}{2}}} \qquad (22)$$

Equations 21 and 22 are the equations we want.

### Input and Output Reflection Coefficients

We now need to derive the expressions for the reflection coefficients looking into each port. Beginning with port 1, we have the circuit shown in Figure 3. From Figure 3 we can see that $a_1$ is the

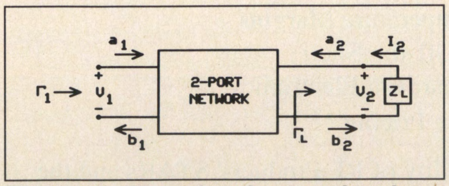


**Figure 3. The circuit showing how the load affects the input reflection coefficient.**

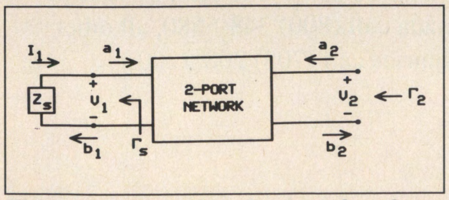


**Figure 4. The circuit showing how the source affects the output reflection coefficient.**

wave incident upon the input and $b_1$ is the wave reflected from the input. Reflection coefficient is defined as the reflected wave divided by the incident wave so we write

$$\Gamma_1 = \frac{b_1}{a_1} \qquad (23)$$

In words, $\Gamma_1$ is the input reflection coefficient, the reflection coefficient looking in port 1 with port 2 terminated in a load impedance $Z_L$.

To derive an expression for $\Gamma_1$ in terms of the S-parameters and the load impedance, $Z_L$, divide equation 13a by $a_1$ to get,

$$\frac{b_1}{a_1} = S_{11} + S_{12}\left[\frac{a_2}{a_1}\right]$$

Substituting equation 23 into this gives,

$$\Gamma_1 = S_{11} + S_{12}\left[\frac{a_2}{a_1}\right] \qquad (24)$$

Solve equations 13b for $a_1$ to get,

$$a_1 = \frac{b_2 - S_{22}a_2}{S_{21}} \qquad (25)$$

and substituting this into equation 24 gives,

$$\Gamma_1 = S_{11} + \frac{S_{12}S_{21}(a_a / b_2)}{1 - S_{22}(a_2 / b_2)} \qquad (26)$$

From Figure 3 we can see that $b_2$ is the wave incident upon $Z_L$ and $a_2$ is the wave reflected from $Z_L$. We can then write,

$$\Gamma_L = \frac{a_2}{b_2} \qquad (27)$$

Using equation 27, equation 26 can be rewritten as,

$$\Gamma_1 = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \qquad (28)$$

We need to find an expression for $a_2/b_2$. To do this we will use the expressions derived in equations 21 and 22 where the subscript j is now 2 for port 2. From the current and voltage defined in Figure 3 we can write,

$$V_2 = -I_2 Z_L \qquad (29)$$

Substituting this into equations 21 and 22 written for port 2, we get,

$$a_2 = \frac{-I_2(Z_L - Z_{02})}{\left[2(Z_{02} + Z_{02}^*)\right]^{\frac{1}{2}}} \qquad (30)$$

$$b_2 = \frac{-I_2(Z_L + Z_{02}^*)}{\left[2(Z_{02} + Z_{02}^*)\right]^{\frac{1}{2}}} \qquad (31)$$

Dividing equation 30 by equation 31 and using equation 27 gives,

$$\Gamma_L = \frac{(Z_L - Z_{02})}{(Z_L + Z_{02}^*)} \qquad (32)$$

Notice in equation 24 that if $a_2 = 0$ then $\Gamma_1 = S_{11}$. Writing $S_{11} = b_1/a_1$ with $a_2 = 0$ is the usual method for defining $S_{11}$. However, it is usually not clear how to make $a_2 = 0$. Simply looking at the circuit gives no clue. If one quickly says

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

as is too often done, the question arises, (or should arise) "What is $Z_0$?" $Z_0$ is not in the circuit and the physical interpretation of $a_2 = 0$ is lost. However, from equation 30 we can see that to make $a_2$ equal to 0, $Z_L$ must equal $Z_{02}$, i.e. the load impedance must be the same as the normalizing impedance. In other words, $\Gamma_L$ is the reflection coefficient relative to the port 2 normalizing impedance.

In a similar fashion, an expression can be obtained for the output reflection coefficient, the reflection coefficient looking in port 2. This circuit is shown in Figure 4. From the circuit we see that,

$$\Gamma_2 = \frac{b_2}{a_2} \tag{33}$$

Solving equation 13b for $b_2/a_2$ and substituting equation 33 gives,

$$\Gamma_2 = S_{21}\left[\frac{a_1}{a_2}\right] + S_{22} \tag{34}$$

Then from equation 13a,

$$a_2 = \frac{b_1 - S_{11}a_1}{S_{12}} \tag{35}$$

Substituting this into equation 34,

$$\Gamma_2 = S_{22} + \frac{S_{12}S_{21}(a_1/b_1)}{1 - S_{11}(a_1/b_1)} \tag{36}$$

From Figure 4 we can see that $b_1$ is the wave incident upon the source and that $a_1$ is the wave reflected from the source. We can then write,

$$\Gamma_s = \frac{a_1}{b_1} \tag{37}$$

and using it in equation 36 we can write,

$$\Gamma_2 = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{22}\Gamma_S} \tag{38}$$

Now we need to find an expression for $a_1/b_1$. To do this we again use the expressions derived in equations 21 and 22. From the circuit in Figure 4, we can write,

$$V_1 = -I_1 Z_S \tag{39}$$

Using this in equations 21 and 22, where now j is 1 for port 1, we have,

$$a_1 = \frac{-I_1(Z_S - Z_{01})}{\left[2(Z_{01} + Z_{01}^*)\right]^{\frac{1}{2}}} \tag{40}$$

$$b_1 = \frac{-I_1(Z_S + Z_{01}^*)}{\left[2(Z_{01} + Z_{01}^*)\right]^{\frac{1}{2}}} \tag{41}$$

Dividing equation 40 by equation 41 and using equation 37 gives,

$$\Gamma_S = \frac{(Z_S - Z_{01})}{(Z_S + Z_{01}^*)} \tag{42}$$

These equations show that if the source impedance is equal to the normalizing impedance, $\Gamma_S = 0$ and $\Gamma_2 = S_{22}$. In other words, $\Gamma_S$ is the reflection coefficient relative to the port 1 normalizing impedance.

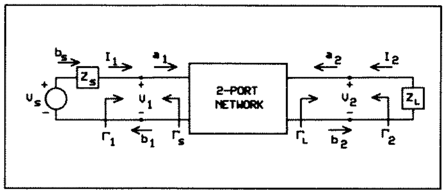


**Figure 5. The complete circuit for which the gain is to be calculated.**

## Gain Equations

We now want to begin developing equations for gain. There are different gains that can be defined but the basic circuit is shown in Figure 5.

## Power Relations

First, we must derive equations for the power into a port. From circuit theory we know that the time average power into port 1 is given by,

$$P_1 = \frac{1}{2}\cdot \mathrm{Re}\{V_1 I_1^*\} \tag{43}$$

To get an expression for this, first subtract equation 22 from equation 21 to get,

$$a_1 - b_1 = I_1 \cdot \left[\frac{(Z_{01} + Z_{01}^*)}{2}\right]^{\frac{1}{2}} \tag{44}$$

Solving this for $I_1$.

$$I_1 = \left[\frac{2}{(Z_{01} + Z_{01}^*)}\right]^{\frac{1}{2}} (a_1 - b_1) \tag{45}$$

Then, adding equations 21 and 22 gives,

$$a_1 + b_1 = \frac{2V_1 + I_1(Z_{01} - Z_{01}^*)}{\left[2(Z_{01} + Z_{01}^*)\right]^{\frac{1}{2}}} \tag{46}$$

Substituting equation 45 into equation 46 gives,

$$a_1 + b_1 = \frac{2V_1 + \left[(a_1 - b_1)\left[\frac{2}{(Z_{01} + Z_{01}^*)}\right]^{\frac{1}{2}}\right](Z_{01} - Z_{01}^*)}{\left[2(Z_{01} + Z_{01}^*)\right]^{\frac{1}{2}}}$$

Solving for $V_1$ gives,

$$V_1 = \left[\frac{2}{(Z_{01} + Z_{01}^*)}\right]^{\frac{1}{2}} (a_1 Z_{01}^* + b_1 Z_{01}) \tag{47}$$

Now, substitute equations 45 and 47 into equation 43 to get,

$$P_1 = \frac{1}{2}\cdot \mathrm{Re}\left[\left[\frac{2}{(Z_{01} + Z_{01}^*)}\right]^{\frac{1}{2}} (a_1 Z_{01}^* + b_1 Z_{01}) \bullet \left[\left[\frac{2}{(Z_{01} + Z_{01}^*)}\right]^{\frac{1}{2}} (a_1 - b_1)\right]^*\right]$$

Simplifying, and using the identities

$$(x + y)^* = x^* + y^*$$

$$(xy)^* = x^* y^*$$

$$(x^{\frac{1}{2}})^* = (x^*)^{\frac{1}{2}}$$

gives,

$$P_1 = \mathrm{Re}\left[\frac{a_1 a_1^* Z_{01}^* - a_1 b_1^* Z_{01}^* + a_1^* b_1 Z_{01} - b_1 b_1^* Z_{01}}{(Z_{01} + Z_{01}^*)}\right]$$

Notice that $a_1 b_1^* Z_{01}^* = (a_1^* b_1 Z_{01})^*$ and that a number subtracted from its conjugate is a purely imaginary number (2). Therefore we can further reduce this equation to,

$$P_1 = \mathrm{Re}\left[\frac{a_1 a_1^* Z_{01}^* - b_1 b_1^* Z_{01}}{(Z_{01} + Z_{01}^*)}\right]$$

Making use of the identity

$$xx^* = |x|^2$$

we can write

$$P_1 = \mathrm{Re}\left[\frac{|a_1|^2 Z_{01}^* - |b_1|^2 Z_{01}}{(Z_{01} + Z_{01}^*)}\right]$$

$$= \mathrm{Re}\left[\frac{|a_1|^2 Z_{01}^*}{(Z_{01} + Z_{01}^*)}\right] - \mathrm{Re}\left[\frac{|b_1|^2 Z_{01}}{(Z_{01} + Z_{01}^*)}\right]$$

Then noting that

$$x + x^* = 2\mathrm{Re}\{x\}$$

we can write,

$$P_1 = \frac{|a_1|^2}{2R_{01}}\cdot \mathrm{Re}\{Z_{01}^*\} - \frac{|b_1|^2}{2R_{01}}\cdot \mathrm{Re}\{Z_{01}\}$$

where $R_{01} = \mathrm{Re}\{Z_{01}\}$. But $\mathrm{Re}\{Z_{01}^*\} = \mathrm{Re}\{Z_{01}\} = R_{01}$ so,

$$P_1 = \frac{1}{2}\cdot\left[|a_1|^2 - |b_1|^2\right] \tag{48}$$

Equation 48 gives the power delivered to a port. It can be described as the difference between the power to the port, $|a_1|^2$, and the power reflected from the port, $|b_1|^2$.

If we were to derive a similar equation for the power from port 2 to the load, the equation, from the circuit in Figure 5, would be written,

$$P_{del} = \frac{1}{2}\cdot \mathrm{Re}\{V_2(-I_2)^*\}$$

Since the only difference is the negative sign associated with $I_2$, it's easy to see that the final result would be,

$$P_{del} = \frac{1}{2}\cdot\left[|b_2|^2 - |a_2|^2\right] \qquad (49)$$

## Transducer Power Gain

The primary gain is called the transducer gain, $G_T$, and is defined as,

$$G_T = \frac{P_{del}}{P_{avs}} \qquad (50)$$

where $P_{del}$ is the power delivered to the load and $P_{avs}$ is the power available from the source. $P_{avs}$ is calculated as if the source was connected to a conjugately matched load. If a source is conjugately matched to a load, it can be shown that the reflection coefficient looking from the source to the load is the conjugate of the reflection coefficient looking from the load to the source. So, for a conjugately matched load, using the notation of Figure 5, $\Gamma_1 = \Gamma_S{}^*$.

*1) Calculation of $P_{avs}$:* To calculate $P_{avs}$, we start with equation 48. From before, we know that,

$$\Gamma_1 = \frac{b_1}{a_1}$$

or

$$b_1 = a_1\Gamma_1 \cdot \qquad (51)$$

From Figure 5 we can write.

$$V_S = I_1 Z_S + V_1$$

Substituting in the expressions for $V_1$ and $I_1$ given in equations 47 and 45 and solving for $a_1$ we have,

$$a_1 = \frac{V_S}{(Z_S + Z_{01}{}^*)}\cdot\left[\frac{Z_{01}+Z_{01}{}^*}{2}\right]^{\frac{1}{2}} + b_1\cdot\left[\frac{Z_S - Z_{01}}{Z_S + Z_{01}{}^*}\right]$$

Making use of equation 42 we can write,

$$a_1 = b_S + b_1\Gamma_S \qquad (52)$$

where $b_S = \dfrac{V_S}{(Z_S + Z_{01}{}^*)}\cdot\left[\dfrac{Z_{01}+Z_{01}{}^*}{2}\right]^{\frac{1}{2}}$.

Equation 52 shows that the $a_1$ consists of a signal from the source, $b_S$, as well as a portion of $b_1$ reflected from the source impedance.

Substituting equation 51 into 52 gives,

$$a_1 = b_S + a_1\Gamma_1\Gamma_S$$

Solving for $a_1$,

$$a_1 = \frac{b_S}{1-\Gamma_S\Gamma_1} \qquad (53)$$

Using equation 51 in equation 48 gives,

$$P_1 = \frac{1}{2}\cdot|a_1|^2(1-|\Gamma_1|^2) \qquad (54)$$

Using equation 53 in 54 gives,

$$P_1 = \frac{1}{2}\cdot\frac{|b_s|^2(1-|\Gamma_1|^2)}{|1-\Gamma_s\Gamma_1|^2}$$

The power available from the source is defined with $\Gamma_1 = \Gamma_S{}^*$. So, substituting this into the expression for P1 gives,

$$P_{avs} = \frac{1}{2}\cdot\frac{|b_s|^2(1-|\Gamma_s|^2)}{\left|1-|\Gamma_s|^2\right|^2}$$

or

$$P_{avs} = \frac{1}{2}\cdot\frac{|b_S|^2}{1-|\Gamma_S|^2} \qquad (55)$$

*2) Calculation of $P_{del}$:* Looking at the load end of the circuit in Figure 5, we can write,

$$\Gamma_L = \frac{a_2}{b_2}$$

or

$$a_2 = b_2\Gamma_L \qquad (56)$$

Substituting this into equation 49 gives,

$$P_{del} = \frac{1}{2}\cdot|b_2|^2(1-|\Gamma_L|^2) \qquad (57)$$

Using equations 55 and 57 in equation 50,

$$G_T = \frac{|b_2|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{|b_S|^2} \qquad (58)$$

Now we need an expression for $b_2/b_S$. Using the approach in Vendelin, Pavio, and Rohde, (3), we use equation 56 in equation 13b to get,

$$b_2 = S_{21}a_1 + S_{22}b_2\Gamma_L$$

Solving for $b_2$ gives,

$$b_2 = \frac{S_{21}a_1}{1-S_{22}\Gamma_L}$$

Dividing by $a_1$ gives,

$$\frac{b_2}{a_1} = \frac{S_{21}}{1-S_{22}\Gamma_L} \qquad (59)$$

From equation 53 we can write,

$$\frac{a_1}{b_S} = \frac{1}{1-\Gamma_s\Gamma_1} \qquad (60)$$

Multiplying equation 59 by 60 will give

$$\frac{b_2}{b_S} = \frac{S_{21}}{(1-\Gamma_S\Gamma_1)(1-S_{22}\Gamma_L)} \qquad (61)$$

and substituting this into equation 58 gives,

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{|1-\Gamma_S\Gamma_1|^2|1-S_{22}\Gamma_L|^2} \qquad (62)$$

This is the first of three expressions for transducer gain.

If we make use of equation 28, we can write,

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{\left|1-\Gamma_S\cdot\left[S_{11}+\dfrac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}\right]\right|^2\cdot|1-S_{22}\Gamma_L|^2}$$

Reducing, we get,

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{|(1-S_{11}\Gamma_S)(1-S_{22}\Gamma_L)-S_{12}S_{21}\Gamma_L\Gamma_S|^2} \qquad (63)$$

This is the second expression for the transducer gain.

To derive a third expression, factor $(1 - S_{11}\Gamma_S)$ out of the denominator and write,

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{\left|(1-S_{11}\Gamma_S)\cdot\left[(1-S_{22}\Gamma_L)-\dfrac{S_{12}S_{21}\Gamma_L\Gamma_S}{(1-S_{11}\Gamma_S)}\right]\right|^2}$$

Further simplification gives,

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{\left|(1-S_{11}\Gamma_S)\cdot\left[1-\Gamma_L\cdot\left[S_{22}+\frac{S_{12}S_{21}\Gamma_S}{(1-S_{11}\Gamma_S)}\right]\right]\right|^2}$$

Then substitute $\Gamma_2$ from equation 38 to get,

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{|1-S_{11}\Gamma_S|^2|1-\Gamma_L\Gamma_2|^2} \quad (64)$$

This is the third form of the transducer gain equation.

### Operating Power Gain

The operating power gain, $G_P$, is defined as the ratio of the power delivered to the load, $P_{del}$, to the power input to the network, $P_{in}$.

$$G_P = \frac{P_{del}}{P_{in}} \quad (65)$$

Equation 57 gives an expression for the power delivered to the load. Equation 54 gives an expression for the power into a port. Rewriting equation 54 but now writing $P_{in}$ for $P_1$, we have,

$$P_{in} = \frac{1}{2}\cdot|a_1|^2(1-|\Gamma_1|^2) \quad (54)$$

Substituting $a_1$ of equation equation 53 into this gives,

$$P_{in} = \frac{1}{2}\cdot\frac{|b_S|^2(1-|\Gamma_1|^2)}{|1-\Gamma_S\Gamma_1|^2} \quad (66)$$

The power delivered to the load, $P_{del}$, is the same as in the previous section, so,

$$P_{del} = \frac{1}{2}\cdot|b_2|^2(1-|\Gamma_L|^2) \quad (57)$$

Substituting equations 57 and 66 into equation 65 gives,

$$G_P = \frac{|b_2|^2(1-|\Gamma_L|^2)(|1-\Gamma_S\Gamma_1|^2)}{|b_S|^2(1-|\Gamma_1|^2)}$$

Then, using $b_2/b_S$ in equation 61

$$G_P = \frac{|S_{21}|^2(1-|\Gamma_L|^2)(|1-\Gamma_S\Gamma_1|^2)}{|1-S_{22}\Gamma_L|^2|1-\Gamma_S\Gamma_1|^2(1-|\Gamma_1|^2)}$$

Simplifying we get,

$$G_P = \frac{|S_{21}|^2(1-|\Gamma_L|^2)}{|1-S_{22}\Gamma_L|^2(1-|\Gamma_1|^2)} \quad (67)$$

Another way to arrive at this result is to substitute $\Gamma_S = \Gamma_1{}^*$ into the transducer gain equation as it is written in equation 62. While this is simpler, a person looses sight of the meaning. $P_{in}$ is the actual power into the network, therefore the substitution $\Gamma_1 = \Gamma_S{}^*$ that was used to write $P_{avs}$ in $G_T$ should not have been made. By using $\Gamma_1{}^*$ in place of $\Gamma_S$ we are essentially reversing the earlier substitution.

### Available Power Gain

In addition to the transducer gain and operating power gain, available power gain is also defined. The available gain, $G_A$, is the ratio of the power available from the network, $P_{avn}$, to the power available from the source, $P_{avs}$.

$$G_A = \frac{P_{avn}}{P_{avs}} \quad (68)$$

The power available to the load is simply the power that would be delivered if the load was conjugately matched to the network, i.e. $\Gamma_L = \Gamma_2{}^*$. To find an expression for it, simply substitute $\Gamma_2{}^*$ for $\Gamma_L$ in equation 64. Reducing we have,

$$G_A = \frac{|S_{21}|^2(1-|\Gamma_S|^2)}{|1-S_{11}\Gamma_S|^2(1-|\Gamma_2|^2)} \quad (69)$$

### Application

A two-port device, when placed in a system with a certain source and load impedance, will produce a certain gain. This gain is, of course, independent of the source and load impedance of the system in which the S-parameters were determined. To show that the equations developed in this paper are correct, the gain for a two-port device will be calculated using S-parameters calculated in two different systems. In one case, S-parameters that were determined in a 50 ohm system (i.e. $Z_{01} = Z_{02} = 50$) will be used. In another case, S-parameters that were determined in a system with a source impedance equal to 10 + j 70 and a load impedance equal to 75 – j 30 (i.e. $Z_{01}$ = 10 + j 70 and $Z_{02}$ = 75 – j 30). Then, in both cases the gain will be calculated for that two-port in a 50 ohm system ($Z_S = Z_L = 50$).

The two-port device is a pad network consisting of ideal resistors as shown in Figure 6. Table 1 shows the parameters used in the different cases and the results of the calculations. The numbers in the parenthesis indicate the equation

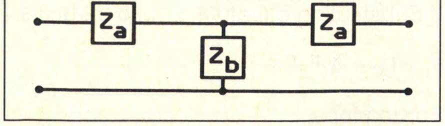


**Figure 6. A simple 10 dB pad in a 50 ohm system. $Z_a$ = 25 ohms, $Z_b$ = 37.5 ohms.**

| | CASE 1 | | CASE 2 | |
|---|---|---|---|---|
| $Z_{01}$<br>$Z_{02}$ | 50 + j 0<br>50 + j 0 | | 10 + j 70<br>75 − j 30 | |
| | Mag | Ang | Mag | Ang |
| $S_{11}$ | 0.0 | 0.0° | 0.868 | 10.6° |
| $S_{12}$ | 0.333 | 0.0° | 0.158 | −34.9° |
| $S_{21}$ | 0.333 | 0.0° | 0.158 | −34.9° |
| $S_{22}$ | 0.0 | 0.0° | 0.235 | −128.9° |
| $Z_S$<br>$Z_L$ | 50 + j 0<br>50 + j 0 | | 50 + j 0<br>50 + j 0 | |
| $\Gamma_S$ (42)<br>$\Gamma_L$ (32) | 0<br>0 | | 0.859 − j 0.165<br>−0.135 + j 0.272 | |
| $G_T$ (63)<br>$G_T$ (dB) | 0.111<br>−9.54 | | 0.111<br>−9.54 | |

**Table 1. The parameters used and calculated to show that $\Gamma_S$ and $\Gamma_L$ are reflection coefficients relative to the normalizing impedance. The gain of the two-port is independent of the system in which the S-parameters were determined.**

used in the calculation. As expected, the gain is the same in both cases.

## Conclusion

This paper presented the derivation of several important equations used in microwave amplifier design using definitions from circuit theory. Equations for the input and output reflection coefficients were derived and with them, expressions for $\Gamma_S$ and $\Gamma_L$. It was shown that $\Gamma_S$ and $\Gamma_L$ are reflection coefficients of the source and load impedances relative to the normalizing impedances. The gain equations were then derived strictly from definitions. It is hoped that this approach more clearly shows the source of the terms in these equations and makes them easier to use. *RF*


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### About the Author


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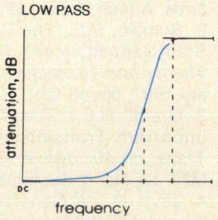


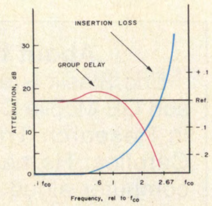


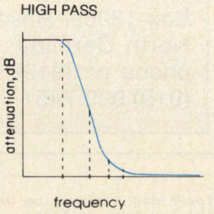


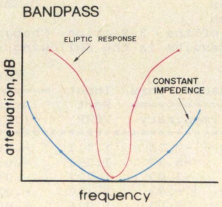


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| ★LP-30 | DC-32 | 47-61 | 61-200 |
| ★LP-50 | DC-48 | 70-90 | 90-200 |
| ★LP-70 | DC-60 | 90-117 | 117-300 |
| ★P-90 | DC-81 | 121-137 | 167-400 |
| ★LP-100 | DC-98 | 146-189 | 189-400 |
| ★LP-150 | DC-140 | 210-300 | 300-600 |
| ★LP-200 | DC-190 | 290-390 | 390-800 |

| Model No. | Passband MHz loss < 1dB | Stopband, MHz loss > 20dB | Stopband, MHz loss > 40dB |
|---|---|---|---|
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| ★LP-300 | DC-270 | 410-550 | 550-1200 |
| ★LP-450 | DC-400 | 580-750 | 750-1800 |
| ★LP-550 | DC-520 | 750-920 | 920-2000 |
| ★LP-600 | DC-680 | 840-1120 | 1120-2000 |
| ★LP-750 | DC-700 | 1000-1300 | 1300-2000 |
| ★LP-800 | DC-720 | 1080-1400 | 1400-2000 |
| ★LP-850 | DC-760 | 1100-1400 | 1400-2000 |
| ★LP-1000 | DC-900 | 1340-1750 | 1750-2000 |
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| SCLF-45 | DC-45 | 70-90 | 90-200 |
| SCLF-135 | DC-135 | 210-300 | 300-600 |
| SCLF-190 | DC-190 | 290-390 | 390-800 |
| SCLF-380 | DC-380 | 580-750 | 750-1800 |
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|---|---|---|---|---|---|---|---|---|
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| ★BLP-117 | DC-65 | 234-312 | 312 | 1.3:1 | 2.4:1 | 0.35 | 1.4 | 1.9 |
| ★BLP-156 | DC-94 | 312-416 | 416 | 0.3:1 | 1.1:1 | 0.3 | 1.1 | 1.5 |
| ★BLP-200 | DC-120 | 400-534 | 534 | 1.6:1 | 1.9:1 | 0.4 | 1.3 | 1.6 |
| ★BLP-300 | DC-180 | 600-801 | 801 | 1.25:1 | 2.2:1 | 0.2 | 0.6 | 0.8 |
| ★BLP-467 | DC-280 | 934-1246 | 1246 | 1.25:1 | 2.2:1 | 0.15 | 0.4 | 0.55 |
| ▲BLP-933 | DC-560 | 1866-2490 | 2490 | 1.3:1 | 2.2:1 | 0.09 | 0.2 | 0.28 |
| ▲BLP-1870 | DC-850 | 3740-6000 | 5000 | 1.45:1 | 2.9:1 | 0.05 | 0.1 | 0.15 |

Price, (1-9 qty), all models: plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE: ▲: -933 and -1870 only with connectors, at additional $2 above other connector models.

### high pass, Plug-in, 27.5 to 2200MHz

| Model No. | Stopband MHz loss < 40dB | Stopband MHz loss < 20dB | Passband, MHz loss < 1dB | VSWR Passband Typ. |
|---|---|---|---|---|
| ★HP-25 | DC-13 | 13-19 | 27.5-200 | 1.8:1 |
| ★HP-50 | DC-20 | 20-26 | 41-200 | 1.5:1 |
| ★HP-100 | DC-40 | 40-55 | 90-400 | 1.8:1 |
| ★HP-150 | DC-70 | 70-95 | 133-600 | 1.8:1 |
| ★HP-175 | DC-70 | 70-105 | 160-800 | 1.5:1 |
| ★HP-200 | DC-90 | 90-116 | 185-800 | 1.6:1 |
| ★HP-250 | DC-100 | 100-150 | 225-1200 | 1.3:1 |
| ★HP-300 | DC-145 | 145-170 | 290-1200 | 1.7:1 |

| Model No. | Stopband MHz loss < 40dB | Stopband MHz loss < 20dB | Passband, MHz loss < 1dB | VSWR Passband Typ. |
|---|---|---|---|---|
| ★HP-400 | DC-210 | 210-290 | 395-1600 | 1.7:1 |
| ★HP-500 | DC-280 | 280-365 | 500-1600 | 1.8:1 |
| ★HP-600 | DC-350 | 350-440 | 600-1600 | 2.0:1 |
| ★HP-700 | DC-400 | 400-520 | 700-1800 | 1.6:1 |
| ★HP-800 | DC-445 | 445-570 | 780-2000 | 2.1:1 |
| ★HP-900 | DC-520 | 520-660 | 910-2100 | 1.8:1 |
| ★HP-1000 | DC-550 | 550-720 | 1000-2200 | 1.9:1 |

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95

### bandpass, Elliptic Response, 10.7 to 70MHz

| Model No. | Center Freq. (MHz) | Passband I.L. 1.5 dB Max. (MHz) | 3 dB Bandwidth Typ. (MHz) | Stopbands I.L. > 20dB at MHz | Stopbands I.L. > 35dB at MHz |
|---|---|---|---|---|---|
| ★BP-10.7 | 10.7 | 9.6-11.5 | 8.9-12.7 | 7.5 & 15 | 0.6 & 50-1000 |
| ★BP-21.4 | 21.4 | 19.2-23.6 | 17.9-25.3 | 15.5 & 29 | 3.0 & 80-1000 |
| ★BP-30 | 30.0 | 27.0-33.0 | 25-35 | 22 & 40 | 3.2 & 99-1000 |
| ★BP-60 | 60.0 | 55.0-67.0 | 49.5-70.5 | 44 & 79 | 4.6 & 190-1000 |
| ★BP-70 | 70.0 | 63.0-77.0 | 68.0-82.0 | 51 & 94 | 6.0 & 193-1000 |

Price, (1-9 qty), all models: plug-in $18.95,
BNC $40.95, SMA $42.95, Type N $43.95

### Constant Impedance, 21.4 to 70MHz

| Model No. | Center Freq. MHz | Passband MHz loss < 1dB | Stopband loss > 20dB at MHz | VSWR 1.3:1 Total Band MHz |
|---|---|---|---|---|
| ★IF-21.4 | 21.4 | 18-25 | 1.3 & 150 | DC-220 |
| ★IF-30 | 30 | 25-35 | 1.9 & 210 | DC-330 |
| ★IF-40 | 42 | 35-49 | 2.6 & 300 | DC-400 |
| ★IF-50 | 50 | 41-58 | 3.1 & 350 | DC-440 |
| ★IF-60 | 60 | 50-70 | 3.8 & 400 | DC-500 |
| ★IF-70 | 70 | 58-82 | 4.4 & 490 | DC-550 |

Price, (1-9 qty), all models: plug-in $14.95,
BNC $36.95, SMA $38.95, Type N $39.95

NOTE: ★**Add Prefix P, B, N, or S for Pin, BNC, N, or SMA connector requirement.**

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# Lowpass Coaxial Filters Synthesized by COAXLPF

*By Eric L. Stasik*
*Ericsson GE Mobile Communications, Inc.*

*This entry to the 1992 RF Design software design contest is called COAXLPF; an acronym for COAXial Low-Pass Filter. As the name suggests, this program designs coaxial low-pass filters. A remarkable feature of this program is its synthesis capability.*

COAXLPF performs the dimensional synthesis of Chebyshev and Butterworth coaxial low-pass filters. The synthesis technique is the well worn transformation of prototype element values into lumped element electrical component values. These lumped element values are then approximated with alternating high and low impedance distributed transmission lines. COAXLPF allows the user to design either a lumped element or distributed element filter so it is useful for those working at frequencies where distributed filters become unwieldy. This feature also helps to illustrate the theory behind the synthesis.

COAXLPF will also perform a frequency analysis on lumped element and distributed filters either synthesized by the program or defined by the user. Clever use of the analysis capability will allow for the frequency analysis of many coaxial discontinuities. I have successfully used this feature to design ultra-low VSWR transitions.

The program is austere, lacking in fancy graphical and colorful presentation. Macht nichts, RF design is an ugly business, not a beauty contest. Time and emphasis has been placed exclusively towards achieving accurate results. Every conceivable discontinuity and frequency effect has been considered. The measured results from dozens of manufactured filters have been used as benchmarks. Agreement between these measurements and theoretical predictions is excellent. (This is not a guarantee, your mileage will depend on your driving habits.)

The program is written in FORTRAN and compiled with Microsoft Corp.'s FORTRAN compiler, version 5.0. The program does not require a math coprocessor but will access one if installed. No special hardware is required, but the software may not run reliably on anything less than a 286 machine. A default data file is read upon invocation so the user can step through the program by just hitting the enter key. This file can be customized by the user to eliminate redundant entries. The references used to create the program code are included below (1 - 5).

## Acknowledgement

The author wishes to acknowledge Mr. Robert M. Garvey and Mr. Mark E. Coles for their able technical support during the development of this program. Mr. Garvey is credited for providing the framework of the ladder analysis executed within the program and Mr. Coles is credited with providing the prototype filter synthesis.

*This program is available on disk from the RF Design Software Service. See page 82 for ordering information.* **RF**

## References

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### About the Author

Eric Stasik is employed with Ericsson GE Mobile Communications, Inc. in Research Triangle Park, North Carolina. He has authored several computer aided engineering design programs, spanning device technology from UHF to 100 GHz. He may be reached at: 1 Triangle Drive, P.O. Box 13969, Research Triangle Park, North Carolina, 27709, or by telephone at: (919) 990-7270 and FAX: (919) 990-7451.

```
COAXLPF (version 1.2)                                    DATE = 14 DEC 1992
                                                         TIME = 10:18:12
      RF/Microwave Computer-Aided Engineering Design Program
                For Coaxial Low-Pass Filters:

        Synthesis of the distributed circuit for your low
            pass filter results in the following data:

            Design Frequencies:      Filter Parameters:
            ===================     ====================
            FC1 = 1000.0000 MHz     Design   = Low Pass
                                    Sections =  3
                                    Response =  Chebyschev
                                    Ripple   =  .100 dB

          Cond. Diameter  Relative             TE11    Peak
Section     (Inches)     Dielectric    Zo     Cutoff   Power    Length
Number     Inner  Outer   Constant   (Ohms)   (GHz)    (kW)    (Inches)
***********************************************************************
Source     .4300 1.0000    1.0000     50.60    5.254  44.484
  1        .7000 1.0000    1.0000     21.39    4.420  49.820     .6636
  2        .1000 1.0000    1.0000    138.06    6.831   6.564     .6923
  3        .7000 1.0000    1.0000     21.39    4.420  49.820     .6636
 Load      .4300 1.0000    1.0000     50.60    5.254  44.484

                                    Total length of filter =  2.0196
```

**Figure 1. Output of synthesis routine.**

```
COAXLPF (version 1.2)                                   DATE = 14 DEC 1992
                                                        TIME = 10:18:12
      RF/Microwave Computer-Aided Engineering Design Program
                For Coaxial Low-Pass Filters:

      Rs = 50.60 Ohms                                 Rl = 50.60 Ohms
      Xs =    .00 Ohms                                Xl =    .00 Ohms
      Sr =   1.0 u-inches                          DLTAN = .0500
     Res =   1.0 u-ohm-cm
     Frequency ANALYSIS of the  3-Section  Chebyschev  Low Pass
    DISTRIBUTED Filter results in the following response profile:

                                                   Thru Path
   Analysis  Input Impedance, Ohms  Input   ===============  Time
   Frequency =====================  Port     Loss    Phase   Delay
     (MHz)      Real    Imaginary   VSWR     (dB)    (Deg)   (nsec)
*****************************************************************
    500.0000   40.69        -7.35   1.311    -.081  -48.81    .273
    600.0000   40.36        -5.42   1.291    -.073  -58.72    .279
    700.0000   41.72        -3.09   1.227    -.048  -68.92    .289
    800.0000   45.32         -.93   1.118    -.016  -79.58    .304
    900.0000   51.91         -.03   1.026    -.004  -90.85    .323
   1000.0000   62.07        -2.96   1.235    -.051 -102.87    .344
   1100.0000   73.70       -14.65   1.559    -.216 -115.60    .362
   1200.0000   76.65       -37.78   2.057    -.557 -128.84    .371
   1300.0000   60.94       -59.02   2.808   -1.115 -142.14    .365
   1400.0000   38.54       -63.99   3.920   -1.891 -154.97    .345
   1500.0000   22.55       -58.59   5.516   -2.850 -166.86    .314
```

**Figure 2. Output of frequency analysis routine.**

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# A Low Cost, Fast RF Switch

*By Ronald "Sam" Zborowski*
*ITS Corporation*

*Some IF processing work requires fast, broadband electronic signal switching. Conventional RF switch designs have limited switching speed and low end bandwidth due to LC circuits used to separate RF and control signal paths. A design which uses the balanced-current cancellation properties of a transmission-line transformer avoids the LC crossover network switching speed limitation and performs well over the inherent bandwidth of the transmission line transformer. The circuit shown in Figure 1 is a SPDT RF switch, but many configurations are possible!*

In the above example the switch bandwidth (–3 dB) is from 1 to 250 MHz. Midband (on) loss is 0.8 dB, midband isolation (off path) is 35 dB, switching time is 20 nsec (with a fast bipolar driver). Component cost in small quantity is about $20 plus circuit board.

## Operation

When the control voltage is positive, diodes CR3 and CR4 are biased on, conducting the RF signal output of T1 to the input of T3 (port B). At that time CR1 and CR2 are reverse biased, isolating port A. On the negative control voltage case, the opposite sets of diodes are biased on and off, directing the RF signal to port A and isolating port B. Note that the DC current flow that biases the active pair of diodes flows through the bifilar-wound transformer in opposite directions in two windings. This property avoids transformer core saturation and inductance effects on switching time and RF signal distortion following the switch transition.

The level shift and bipolar driver circuit that was used with this switch is shown in Figure 2. This is one example of many possible ways to obtain the ±2V control signal from logic-level signals.

## Driver Operation

The 82 ohm and 1k ohm resistors together form a net 75 ohm termination to facilitate driving from a remote video drive pulse. The 1N914 diodes are biased on by the 1k ohm resistor to –12 V to serve as the level shifting element to move the +4V / 0V to nominally +2V / –2V at the voltage follower to drive the RF switch. The 0.01 uF capacitor across the diodes prevents rectification of the drive pulse by the diodes.

The design presented above flows from a need to build an IF gain changing circuit that is stable with time and temperature. It includes two of the RF switches as described, connected back-to-back with variable attenuators between them to achieve two independent gain settings with essentially the same time delay in each path. All of the above is incorporated as part of the IF signal processing circuitry of a UHF TV klystron transmitter system. *RF*

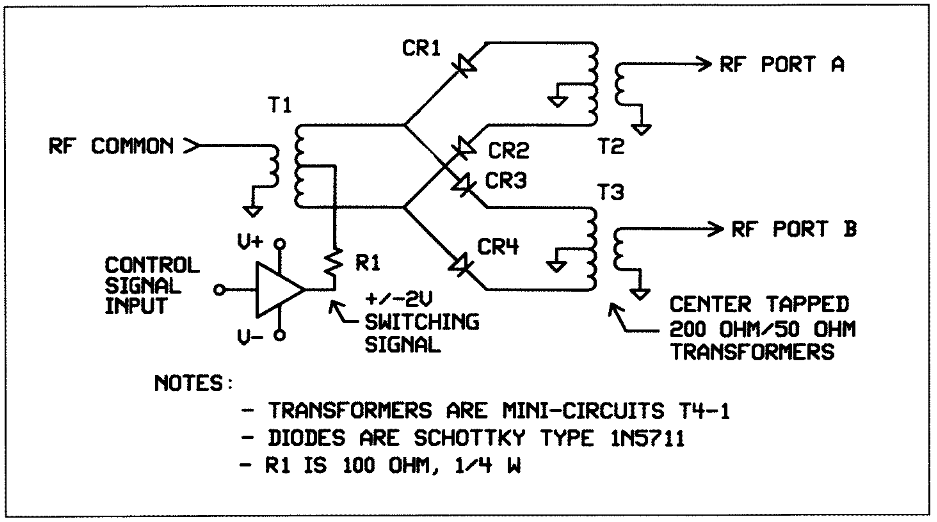


**Figure 1. Schematic of the RF switch.**

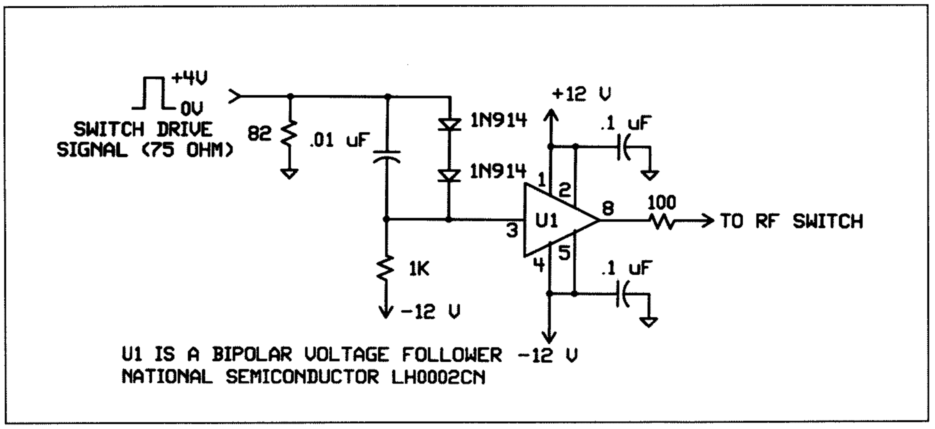


**Figure 2. Schematic of the switch driver circuit.**

### About the Author

Ronald "Sam" Zborowski is Vice-President of Engineering and one of the pricipals of ITS Corporation. He has a BSEE from the University of Pittsburgh and MSE from Pennsylvania State. He can be reached at 375 Valley Brook Rd., McMurray, PA 15317, or by phone at (412) 941-1500.


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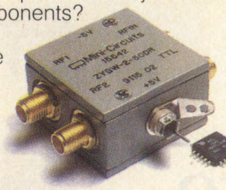

| SPECIFICATIONS (typ) | Absorptive SPDT YSWA-2-50DR ZYSWA-2-50DR | | | Reflective SPDT YSW-2-50DR ZYSW-2-50DR | | |
|---|---|---|---|---|---|---|
| Frequency (MHz) | dc-500 | 500-2000 | 2000-5000 | dc-500 | 500-2000 | 2000-5000 |
| Ins. Loss (dB) | 1.1 | 1.4 | 1.9 | 0.9 | 1.3 | 1.4 |
| Isolation (dB) | 42 | 31 | 20 | 50 | 40 | 28 |
| 1dB Comp. (dBm) | 18 | 20 | 22.5 | 20 | 20 | 24 |
| RF Input (max dBm) | — | 20 | — | 22 | 22 | 26 |
| VSWR "on" | 1.25 | 1.35 | 1.5 | 1.4 | 1.4 | 1.4 |
| Video Bkthru (mV,p/p) | 30 | 30 | 30 | 30 | 30 | 30 |
| Sw. Spd. (nsec) | 3 | 3 | 3 | 3 | 3 | 3 |
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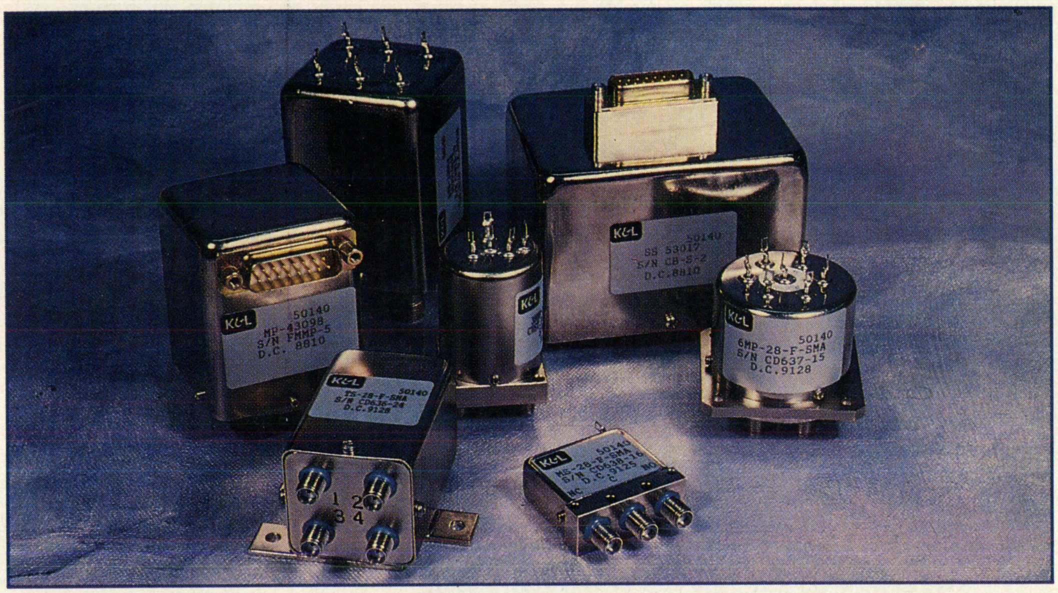

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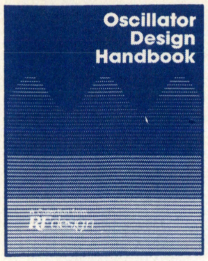


## 1. Oscillator Design Handbook

You can benefit from the contributions of more than 20 top RF engineers as they share their expertise on oscillator theory, design and applications. Analytical topics and practical circuits are presented for crystal oscillators, LC circuits and VCOs. Along with traditional designs, this book includes some unique approaches not found anywhere else.

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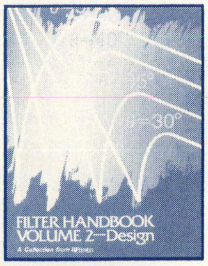


## 2. Filter Handbook: Volume 1 — Applications

The best practical filter circuits from *RF Design* are collected in this book, allowing you to see how the best engineers solved their design problems. Essential information on active, passive, lumped element, microstrip, helical and SAW filters will help make your filter design tasks easier.

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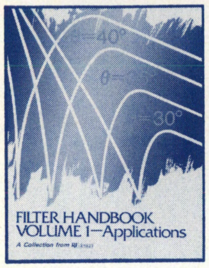


## 3. Filter Handbook: Volume 2 — Design

Do you need to brush up on filter theory and analysis? This book offers fundamental and advanced material on classic Butterworth, Chebyshev and elliptic filters, plus notes on filter implementation, including filter performance with real, not ideal, components. Another highlight is a tutorial series on SAW filter basics.

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## 4. RFI/EMI/EMC: A Designer's Handbook

The best design-for-compliance articles from *RF Design* and *EMC Test & Design* are collected in this practical handbook. Circuit board design, Part 15 techniques, ESD protection, filtering, bypassing and trouble-shooting are among the featured topics. Notes on regulations and test methods are also included, making this a well-rounded collection of EMC and ESD engineering techniques.

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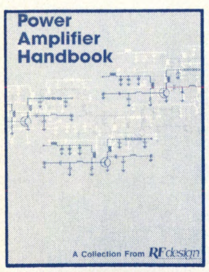


## 5. Power Amplifier Handbook

This book is loaded with practical circuits for power amplifiers operating from HF through L-band, from a few watts to over a kilowatt, with clear explanations of how these circuits were designed. Articles on high power couplers, combiners, biasing techniques and VSWR protection will help simplify the design of your next power amplifier system.

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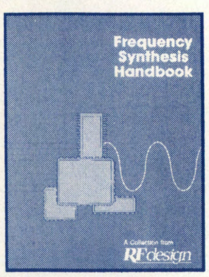


## 6. Frequency Synthesis Handbook

Phase locked loops and direct digital synthesis are the main focus of this handbook, with articles ranging from Andy Przedpelski's ''PLL Primer'' series to advanced analytical techniques. Theoretical material is complemented by practical circuits and application notes on some of the latest synthesizer products.

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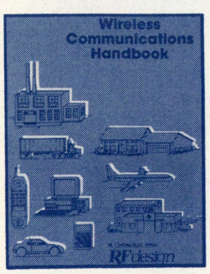


## 7. Wireless Communications Handbook

Engineering methods for new wireless applications are highlighted in this collection from *RF Design*. Topics include spread spectrum systems, Part 15 devices, digital modulation, demodulation, transmission, reception and signal propagation. A special feature is a repeat of our popular tutorial series on Complex Modulation.

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**By Jerry Sevick**


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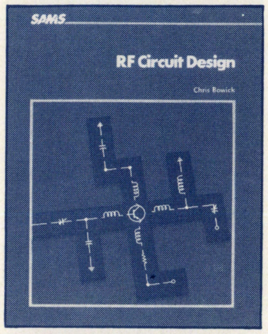
ISBN 0-87259-296-0 American Radio Relay League **$20.00**

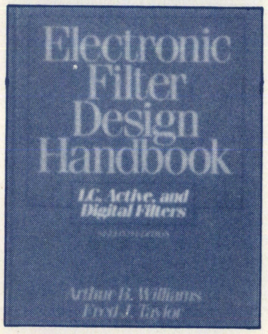
## 9. RF Circuit Design
**By Chris Bowick**




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.

ISBN 0-672-21868-2 Sams 176 pages **$24.00**

## 10. Electronic Filter Design Handbook
**By Arthur B. Williams and Fred J. Taylor**

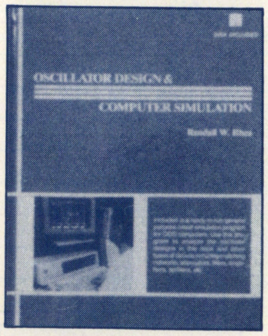



A modern classic for active, passive and digital filter design, this book is an indispensible engineering reference. Basic information is provided on filter types, passband and stopband characteristics, mathematical analysis and various topologies. Extensive tables of filter parameters are included for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

ISBN 0-07-070434-1 McGraw-Hill **$68.00**

TECHNICAL BOOKS

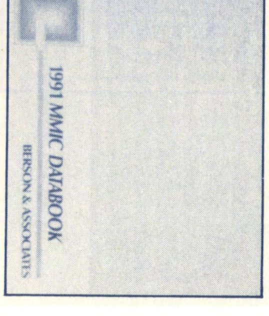
## 11. Oscillator Design and Computer Simulation
**By Randall W. Rhea**




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.

ISBN 0-13-642513-5 Prentice Hall 260 pages **$63.00**

## 12. MMIC Databook




The 1991 MMIC Databook, written with the help of some of the MMIC Industries' most active players, effectively converts "MMIC know-nothings" into "MMIC gurus," and as rapidly as possible. The Databook contains performance graphs based on manufacturers' typical ratings; charts detailing product availability; key contacts for standard, semi-custom and custom foundry information; and foundry data in both datasheet and summary spreadsheet format.

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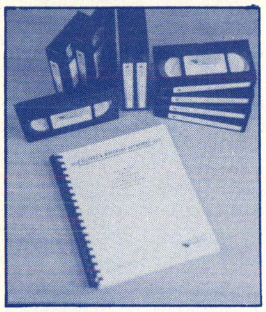

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# RF Expo West Technical Program Abstracts

## Wednesday, March 17 8:30-11:30 a.m.

### Session A-1: Cellular Radio

**A Single Chip Radio Transceiver for 900 MHz GSM Handsets, Mobile Terminals and Base Stations**
***Jan Sevenhans, J. Dulongpont, D. Rabaey, A. Vanwelsnaers, M. Van Paemel, Alcatel Bell***
A single chip solution for a GSM 900 MHz RF TX quadrature modulator and RX quadrature demodulator was designed and fabricated in a 9 GHz silicon bipolar technology. The transmit modulator is a direct upconverter and the receiver demodulator operates in a zero IF architecture. This device is one of the key components of a pocket size hand portable phoneset for application in the GSM pan European digital mobile radio system.

**Cellular Repeaters and Cellular Microcells**
***Ken Anderson, Peninsula Engineering Group, Inc.***
The application of cellular repeaters and/or cellular microcells. These applications are a growing market as cellular networks mature. PCN is a threshold market and the problem with PCNs are similar to the problems with cellular.

**Use of Sigma-Delta Technology to Ease Adjacent Channel Rejection for Mobile Radio Systems**
***Denis O'Mahony, Analog Devices***
In the design of a mobile radio receiver, the filtering of the adjacent channels is an important factor in achieving good receiver performance. Traditionally the use of SAW or Ceramic RF and IF filters are used to perform the necessary filtering. This paper investigates the role of Sigma-Delta technology, examining how this relatively new technology can be applied to filtering requirements. Sigma-Delta ADC's combine the principle of high oversampling and noise shaping with on-board digital filtering to provide an ADC with built-in filtering properties. The paper will also look at the problems and tradeoffs associated with this alternative.

### Session A-2: Antennas and Transmission Lines

**Antenna Field Design for (VHF/UHF) Hand Held and Body Worn Applications**
***Ian Dilworth, University of Essex***
The interaction of the human body with electromagnetic waves depends on frequency and proximity of the antenna to the body. At the VHF and UHF frequencies preferred for PCN applications interaction is relatively strong, for example more than 20 dB variation of field intensity can occur for small movements between the radiator and the body. The radiated E and H field respond differently in loose proximity to the body and this effect is exploited to produce an antenna.

**Ferrite Rod Antenna for AM Broadcast Receivers**
***F. David Harris, Western Washington University***
Except for automotive radios, the ferrite rod antenna is an integral part of almost every AM broadcast radio receiver manufactured since the 1950s. With the recent expansion of the AM broadcast band to 1705 kHz, a close look at the performance of ferrite rod antennas is in order. Using standard rod materials and sizes from a major OEM supplier, and using litz wire and magnet wire for comparison purposes, antenna performance data is gathered and presented for frequencies spanning the entire AM broadcast band, along with implications for radio design.

**New Generation of RF Transmission Line**
***Bernhart A. Gebs, Cooper Industries, Belden Division***
This paper will present the case history of a development program that resulted in a unique RF transmission line. The goal of the development team was to replicate all of the desirable features of semi-rigid cable into a cable that is mechanically formable, hand bendable, and flexible. The paper deals with the mechanical and RF considerations, the design of the complete cable and the issues of quantifying the cable's RF performance requirements, and the design and manufacturing concerns with delivering a 3 sigma product.

### Session A-3: Circuit Analysis Tutorial - Part 1

**Driving Point Impedance Circuit Analysis Technique**
***Kurt Wessendorf, Sandia National Laboratories***
Driving Point Impedance (DPI) is a straightforward technique for rapidly analyzing analog electronic circuitry. DPI greatly reduces the complexity of classical analysis techniques like mesh and nodal analysis in determining circuit impedances and gain transfer functions. Much of the simplicity stems from the use of DPI models for the transistor which allow circuit analysis without having to reduce the transistor into a multi-element model. This one day tutorial will introduce DPI circuit analysis by reviewing and analyzing several classic amplifier circuits. Derivations and practical examples will be discussed for common non-feedback, tuned, and feedback amplifiers including a high frequency 50 ohm gain block.

## Wednesday, March 17 1:30-4:30 p.m.

### Session B-1: Power Amplifier Design

**Use EMTP (ATP) to Understand RF Power Amplifiers**
***Arturo Mediano Heredia, University of Zaragoza***
The use of circuit simulation programs in power electronic design is growing day by day. The circuit-oriented simulator used is the Electro Magnetic Transients Program (EMTP) version called ATP. This software is primarily a simulation program for the electric power industry, predicting variables of interest within electric power networks. It can also be used like an electronics power circuits simulator. In this example a quasi-E topology is employed, and an EMTP (ATP) model is designed for the circuit and for the control system. The circuit parameters are varied in order to study the stress in the device when turned on and off.

**Tradeoffs in Practical Design of Class E High-Efficiency RF Power Amplifiers**
***Nathan Sokal, Laszlo Drimusz and Istvan Novak, Design Automation***
A Class E RF power amplifier would have 100 percent efficiency at any output power and any frequency if it could use ideal components. This paper will explain how efficiency varies with each circuit parameter, and will give a transistor figures of merit which quantify transistor power losses in terms of combinations of transistor parameters. With that background established, the paper will discuss the practical tradeoffs in optimizing a design which uses nonideal (real) components.

**1 kW PWM RF Power Generator for the HF Band**
***George J. Krausse, Directed Energy, Inc.***
The DE-Series of power MOSFET makes it feasible to generate over a kilowatt of power in the HF band, using two devices in push-pull configuration with a conversion efficiency of greater than 85 percent. Designs based on these devices can achieve $0.25 to $0.35 per watt costs (depending on gate drive topology). A class D push pull amplifier is described which delivers over 1 kW in the 13.56 MHz ISM band.

## Session B-2: RF Systems

### Wireless Transmission Using Infrared Techniques
***Robert Rennie, Canadian Marconi Company***

As the number of wireless systems that are developed increases, the spectrum congestion problem demands a more sophisticated receiver design and modulation techniques. Spread spectrum systems, for example, are on the rise. However, transmission in the infrared spectrum provides an attractive alternative. This paper describes an experimental line-of-sight infrared audio link. The system uses FM modulation to transmit an audio signal from an audio component to a remote loudspeaker. The system has also been tested using FSK modulation of NRZ coded data.

### A Medical Radio Frequency Network for the Aged
***Jon W. Swanberg, IEEE***

Aged home-health care data monitoring and transmission will become a critical issue in the near future. The care, treatment and regimen of the aged population will overload the medical community with excessive monitored patient data. To accommodate this data stream, a specific medical RF network would allow a normalized transmission environment for medical data and voice channels. This network might have inception as a IEEE standard, much as the proposed Medical Information Bus.

### Miniaturization Techniques in Receiver Design
***Charles E. Dexter, Watkins-Johnson Company***

A multi conversion superheterodyne receiver is difficult to miniaturize because of electronic component size vs. performance, signal isolation, heat dissipation and costs. The paper concentrates on the "How To" part of miniature RF design that can be useful for a broad range of applications. The RF solutions talked about will be: surface mount technology, signal isolation, heat conduction and dissipation, new circuit designs and future miniaturization techniques.

## B-3: Circuit Analysis Tutorial - Part 2

### Driving Point Impedance Circuit Analysis Technique
***Kurt Wessendorf, Sandia National Laboratories***

(Continued from Session A-3)

# Thursday, March 18 8:30-11:30 a.m.

## Session C-1: HF/VHF Power Amplifiers

### HF Power Amplifier Operates in Both Class B and Class D
***Fred Raab and Daniel Rupp, Green Mountain Radio Research Company***

This transformer-coupled MOSFET power amplifier operates over the HF and lower-VHF bands. Through changes in bias and drive, it can operate in either class B for large bandwidth or class D for high efficiency and maximum output power. The RF-power chain consists of a predriver, driver and a final amplifier which employs a pair of MRF148 MOSFETs connected in push-pull. Operation from a 50-V power supply allows a 100-W output during class-D operation.

### Class-E Power Amplifier Delivers 24 W at 27 MHz at 89-92 percent Efficiency
***Nathan Sokal and Ka-Lon Chu, Design Automation***

Switching-mode RF power amplifiers (e.g. Class E and voltage switching class D) provide significantly higher efficiency than that of Class B and C amplifiers. A Class E power amplifier uses a single power transistor, in contrast with the class D, which uses two or four transistors in a half-bridge or full-bridge topology. This paper reports on a single-transistor class E power amplifier which uses a low cost ($1.05 in 100s) IRF520 MOSFET to deliver 24 W at 27 MHz.

### A New Method of Input/Output Impedance Measurement for Class C/D Amplifiers Using a Spectrum Analyzer
***Climerio dos Santos Viera, Telebras Research and Development Center***

The development of this method was motivated by the need of projecting non-linear amplifiers in the VHF band, to easily and quickly calculate the input/output impedance matching. The method does not need expensive equipment such as network analyzers and narrow band filters. This method allows the impedance measurement of the transistor in ten minutes and it can be repeated any time on the bench, needing just a Smith Chart (graphic mode) or a handheld calculator (analytic mode). A computer program has been written (BASIC), and the method has been used in the design of several class D amplifiers in the 40 MHz to 200 MHz.

## Session C-2: Oscillators

### Modulating SAW Oscillators
***Boni Angelo, REDOX***

Although SAW devices are publicized as easy to modulate devices, in fact it can be difficult to obtain good performance, both in FM or pulse modulation, and some types cannot be reliably modulated at all. The author's experience in the development of consumer circuits in the 300 to 500 MHz frequency range is presented for FM modulation, linear FM modulation, digital FM modulation and pulse modulation, with additional notes on manufacturing a suitable receiver.

### Practical Methods of Computer Simulation of Very High Q Oscillators Using SPICE and LIBRA
***Jeffrey Pawlan, Pawlan Communications***

The computer simulation of low and moderate Q oscillators is relatively easy. However, when the Q is very high, such as when a crystal, SAW or dielectric resonator is used, the engineer finds that computer simulation is difficult and may take many hours to run. This two-hour presentation assumes that the reader already knows how to design and analyze oscillators, but wants to learn how to use computer aided engineering software to verify the design and optimize the performance. The correct method of using a SPICE simulator is presented. Next, the EEsof approach by using LIBRA will be shown. Then more sophisticated methods of using SPICE will be presented. All presented simulation methods will be compared.

## Session C-3: Electromagnetic Modeling

### Simplified Calculation of Planar Antenna Field Perturbation by Integrated Active Feed
***Ian Dilworth, Tito Khan, University of Essex***

This is a relatively simple theoretical modeling technique for quantifying the fields around mismatches and discontinuities in transmission lines connected to 3-port devices whose S parameters can be measured or are known. The paper illustrates calculations for the fields expected around circuits employing active devices and comparisons with rigorous EM calculations show good agreement. In RF applications a powerful role for the tool is the parametric investigation of the third port "S31, S32 and S33," i.e., the ground connection, where the effects of the parasitics associated with the ground route (or VIA) can be investigated through simulation.

# Thursday, March 18 1:30-4:30 p.m.

## Session D-1: UHF and L-Band Power Amplifiers

### Transistors for High Power Linear Amplifiers in the 1.5-2.0 GHz
***David M. Boylan, Gerard David, Jean Pierre Manhout, Philips Semiconductors***

New applications need power amplifiers working in the 1.5 to 2.0 GHz range: personal communication system, digital european cordless telephone, digital audio broadcasting, and mobile satcom. High power levels together with the required linearity lead to choice of class AB transistors. Philips Components has developed for these applications a full range of transistors, mainly characterized under Vcc = 24V at 1.8 GHz. All these transistors are single ended devices, using flat-pack headers and automatic bonding process suited for large volume production.

### 100-450 MHz 250 Watt Power Amplifier
***John de Blok, Philips Semiconductors***

In aviation applications, two frequency ranges are very important; 108-144 MHz and 225-450 MHz. In order to cover these frequency ranges with one amplifier, one is forced to use wideband transistors,

## Improve Your RF Engineering Skills With Special Courses

### Filter and Matching Network Design: L-C and Distributed Circuits — HF to Microwave

*(Tuesday, March 16)* — This course is designed for the practical engineer, packing a wealth of useful information on these passive RF circuits into 8 hours of instruction. Engineers of all levels of experience will benefit from a review of fundamental information on filter response and classic topologies, followed by design methods for implementing these designs. Lumped-element (L-C) and distributed (stripline and microstrip) components are presented for filters and matching networks operating from low RF to microwave frequencies. Key performance parameters of group delay and phase characteristics are covered, as are techniques for implementing design using computer-aided synthesis and analysis. The instructor is Randy Rhea, President of Eagleware.

### Introduction to RF Circuit Design — Part I: Fundamental Concepts

*(Wednesday, March 17)* — New engineers and "old hands" can benefit from this new course, which provides instruction in the basic principles of RF circuits and systems. Part I presents the fundamental concepts of RF systems, components, transmission lines and impedance matching. RF system concepts of gain, bandwidth, linear and nonlinear circuits and amplifier classes are followed by the network concepts of resonance, bandwidth, Q and maximum power transfer. Practical components, their models, characterization and specification are then discussed. Transmission line theory, traveling wave behavior, the Smith chart, and practical realization of transmission lines are discussed, covering the concepts of impedance, loss, dispersion and physical construction. The final topic is impedance transformation networks, including lossless and lossy elements, graphical and analytical design, phase shift and bandwidth considerations, balanced networks and attenuators. Instructors are Dr. Robert K Feeney and Dr. David R. Hertling of the Georgia Tech School of Electrical Engineering.

### Introduction to RF Circuit Design — Part II: Active Circuit Design

*(Thursday, March 18)* — Active circuits are covered in this course, from basic models to advanced computer-aided design methods, from simple two-port concepts to large signal amplifiers and oscillator design. The two-port model includes definition of power gain, noise, cascaded elements and terminations. Scattering parameters used to describe two-ports are introduced, including their definition and physical meanings. Computer-aided design and analysis using SPICE and commercial software is described, noting models, analysis techniques and optimization methods. S parameter based design of amplifiers is presented, followed by large signal amplifier design and the large signal concepts of dynamic range, intermodulation distortion and load pull. Non-reciprocal networks (couplers, combiners and hybrids) are introduced, and the course concludes with oscillator design fundamentals. Instructors are Robert K. Feeney and David R. Hertling.

### Oscillator Design Principles

*(Friday, March 19)* — Learn the fundamentals of oscillator design. Instead of a series of obscure equations for various configurations, this course presents basic concepts that can be applied to oscillator design through a unified approach. Students learn how to evaluate oscillator designs accurately. L-C, distributed element, SAW and crystal oscillators are all included. Also considered are output level, starting time, harmonic levels and phase noise performance. The instructor is Randy Rhea.

which are designed with the output capacitance restricted to the utmost minimum. This paper describes the design, construction and performance of a push-pull two octave broadband amplifier, suitable to generate more than 250 Watts of output power in the frequency range 100-450 MHz, using two Philips MOSFET devices.

### Session D-2: Digital Transmission Systems

#### The Development of X-Band SSPAs and QPSK Modulators for Remote Sensing Satellite Applications

***Brent Stoute, Spar Aerospace***

Image data generated by remote sensing satellites greatly enhances the scientific community's understanding of earth sciences. Spar Aerospace Limited is the prime contractor for the Radarsat Remote Sensing Satellite program which is funded by the Canadian Government, scheduled for launch in early 1994. In remote sensing satellites, instrument data collected by the satellite radar system is transmitted to ground (via down-link) by means of BPSK, UQPSK or QPSK modulation of an X-band carrier in the 8.0 to 8.5 GHz frequency band. This paper will present the SSPA and QPSK modulator development activities conducted at Spar Aerospace for future space programs.

#### Simulating Common Impairments Found in RF Digital Communications Systems

***Charles Plott, Hewlett-Packard Company***

The RF portion of digital radio systems can suffer from a number of real world impairments. Imperfections in the modulator/demodulator, nonlinearities in the transmitter power amplifier, intersymbol interference contributed by non-ideal filtering, and flat- or multipath fading can all adversely affect the bit error rate (BER). This paper steps through a realistic design of pi/4 DQPSK transmitter/receiver pair using the latest software technology to qualitatively and quantitatively predict how the radio's performance withstands different imperfections.

### Session D-3: Component Applications

#### RF Circuits - The ASIC Way

***Paul Paddan, Walmsley Microsystems Ltd.***

This paper presents the development of a mixed analog/digital array, the BTA 496. The array comprises 488 uncommitted npn transistors in cell structures of 1 mA, 5 mA, and 10 mA devices and 8 pnp transistors. Design and layout can be done on low cost PC based platforms. This brings down the economic breakeven point for ASIC design to volumes as low as 1000 devices per annum. Practical circuits developed on the BTA 496 include: 1.2 GHz BW amplifier, 900 MHz oscillator, 100 MHz VCXO with divider cell, fiber optic transimpedance amplifier, LED/laser drivers at 2.4 GHz and 500 MHz graphics shift registers.

#### Tiny Transceiver, Wireless Microphone, and Superhet Receiver

***John Horvath, Minaret Radio***

Several practical miniature circuits will be demonstrated: A "Dick Tracy" type wrist watch transmitter/receiver for low power operation (3 volt), a wireless microphone for 3 V operation, a 600 MHz transmitter for model airplane tracking and a superhet receiver that is tunable with signal strength and audio indicator plus DF antenna, for applications at 200 MHz to 1 GHz.

## Friday, March 19
## 8:30-11:30 a.m.

### Session E-1: Personal Communications

#### A Fully Integrated Modulator/Demodulator for DECT

### Daniel E. Fague, National Semiconductor

A practical, integrated modulator/ demodulator (modem) for the Digital European Cordless Telecommunications (DECT) system is discussed. The modem consists of a quadrature modulator, a frequency discriminator, and a digital baseband chip. The three chip modem is discussed in detail, and is also shown to be part of a complete radio front end for DECT. Simulated eye diagrams and transmit spectra as well as measured eyes, spectra, and bit error rate (BER) curves are presented.

### Vector Modulator ICs for use in Wireless Communication

*Jim Wholey, Hewlett-Packard Company*

This paper discusses two high speed silicon bipolar integrated circuits for use as vector modulators in wireless communication systems. The first modulator provides for direct conversion from baseband to 800-1000 MHz, while the second upconverts baseband signals to 45-200 MHz for use in dual stage upconversion systems. The paper presents a brief overview of the high speed bipolar IC process used in this development and discusses the circuit topology used in the vector modulators. Fundamental measurements such as power levels, mapping errors, and distortion analysis will also be presented.

## Session E-2: Frequency Synthesis and Phase Detection

### Wideband Synthesizer Design

*Steve Anderson, QUALCOMM*

### Operating Principle and Applications of a Frequency Independent Phase Tracking Circuit

*Glen A. Myers, Kintel Technologies*

The paper considers the realization of a novel circuit operating at IF. The circuit provides a bandpass output equal in frequency to that of the input and with a phase offset that does not vary with frequency. The phase offset can be set with an external DC voltage to any value over a 360 degree range. In addition to traditional phase-locked loop applications, practical uses of this circuit include power multiplexing (multiple spectrum reuse), direction finding and beam forming. The paper treats the operation of this patented circuit in several of these applications.

## Session E-3: Computer Methods

### Knowledge Systems: The Fast Track to RF Design and Testing Solutions

*Andrew L. Michuda, Teltech Resource Network Corporation*

Within the framework of the recent knowledge explosion, today's RF engineers are required to make more decisions in less time and on more issues than ever before. Whether the questions have to do with product design, component selection or test methods and standards, the right answers must be found quickly. RF professionals can use modem-equipped PCs and telephones to quickly locate verified sources of materials, components, equipment and services; pick the brains of leading experts; and search the libraries of the world for technical information. The paper will include specific case histories illustrating the impact of knowledge systems use.

### Nonlinear Noise Analysis for Wireless Communication Systems

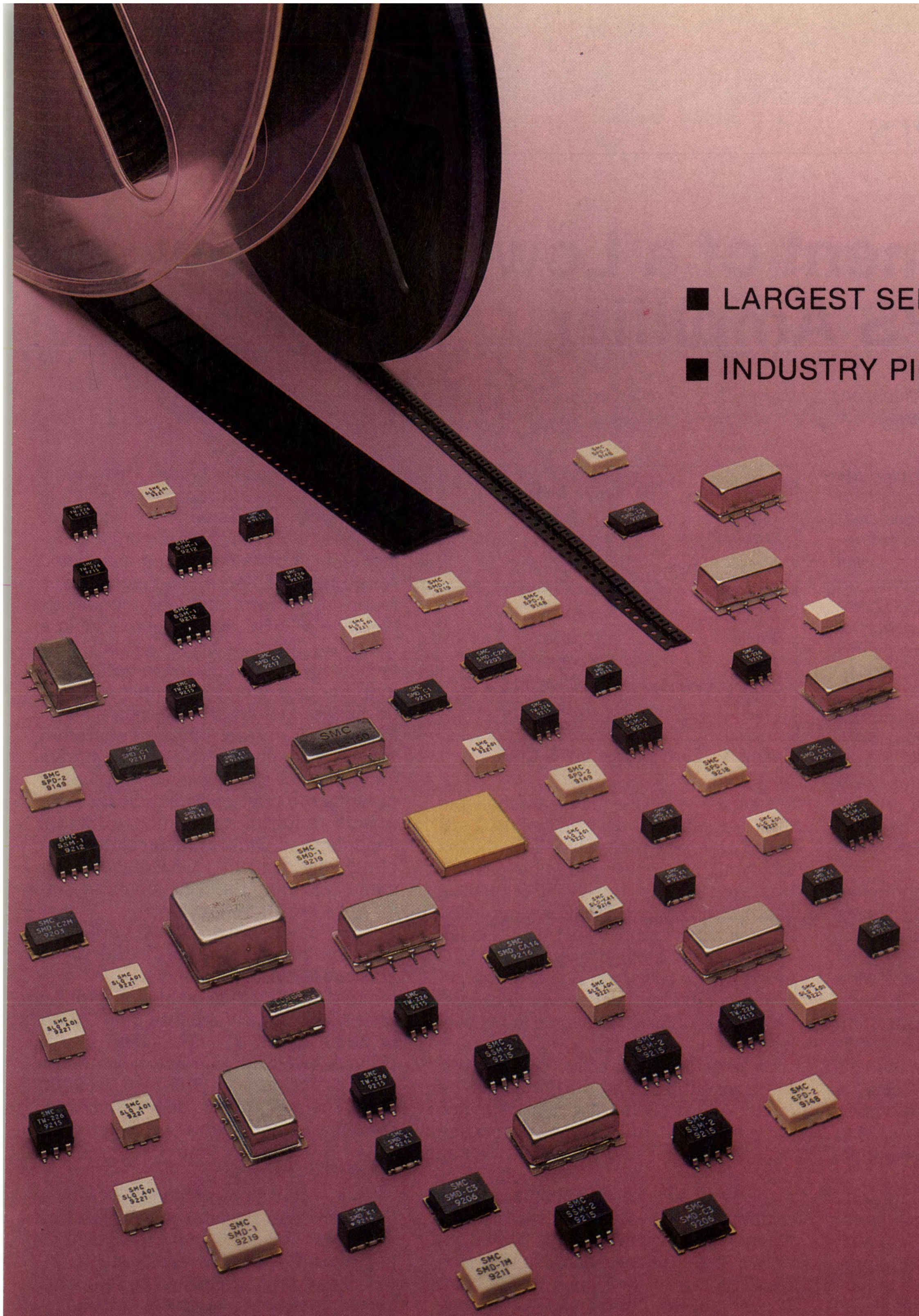
*Chao-Ren Chang and Jason Gerber, Compact Software*

Noise performance is an important determining factor of receiver system sensitivity and dynamic range. Linear simulators are ideal to predict noise performance using arbitrary combinations of active and passive elements. However, in the practical world, most microwave circuits need to be analyzed using nonlinear methods since operating conditions may approach the saturated mode. This paper illustrates how the noise performance of mixer and amplifier circuits is simulated in Microwave Harmonica.

### The Design of Software for Automated RF Testing

*Neal Silence, Consultant*

The design and control of software for RF automated testing can be disappointing from cost and performance standpoints. This paper presents a method for the development of design and control systems for small to medium sized test projects, with an example of the design of a program to conduct a spur search with a spectrum analyzer.



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# Development of a Low Cost, Low Noise GPS Amplifier

*By Paul J. Schwab and Joan Williamson*
*M/A-COM Inc., IC Design Center*

*Glass Microwave Integrated Circuits (GMIC) offer a unique combination of performance and cost advantages over conventional circuit design and manufacturing. Batch processing offers a cost effective manufacturing approach to high volume circuit production. These advantages are utilized in the design of a low cost, low noise amplifier for GPS (global positioning system) applications.*

A new and novel microwave circuit fabrication technology based on a well known dielectric (basically glass) has been exploited to produce very manufacturable, cost effective, and at the same time, high performance low noise amplifiers suitable for GPS applications.

There are a number of technologies currently available such as Si, GaAs, and various other hybrid approaches for manufacturing low noise amplifiers. One usually finds, though, that when a wide range of criteria are considered, as is usually the case in developing a commercial application, each approach carries some penalty and usually there is only one obvious choice. In the case of the GPS amplifier that will be described here, performance, manufacturing, and cost had approximately equal importance.

Due to the relatively small antenna sizes contemplated, (especially for mobile and hand-held applications), and the relatively noisy signal conditions, GPS receiver LNA noise figures of typically less than 2 dB are desirable. Even this leaves very little margin for the input filter. Noise figures of close to 1 dB would be more preferable. This means the circuit elements need to be very low loss, and a good quality FET must be used. Even though MMIC seems to be an obvious first choice, a 1 dB noise figure is just below the typical MMIC foundry capabilities. In addition, the circuit prices do not come down to acceptable levels unless the volume is run very high.

| | |
|---|---|
| Frequency | 1575 MHz |
| Gain | 26 dB (min.) |
| VSWR | 1.5:1 (typ.) |
| Noise Figure | 0.8 dB (typ.) |
| Current | 25mA (max.) |
| DC Voltage | 3.5-6V |
| P1db | +10 dBm |
| IP3 | +20 dbm |
| Size (LxWxH) | 0.135"x0.115"x 0.026" |

**Table 1. Two stage GPS LNA**

By separating the active components, more optimal combinations can be found. After all, it is quite possible to find very low cost but high performance commercial FETs on the market. Not as application specific as a circuit, FETs are produced in huge quantities quite economically. Also, since the discrete technologies are generally a generation or two ahead of the monolithic technologies, as far as the active devices are concerned, the performance of the discrete FET is usually better. This seems to suggest a hybrid approach which is notoriously expensive to manufacture because of extensive assembly requirements, high parts count, and associated reliability problems. The solution is found in the form of a half-monolithic, half hybrid GMIC approach, that is, in a way, the "best of both worlds". This technology has been used to design and manufacture the GPS amplifier to be described. In addition to having state of the art gain, noise figure, temperature stability, linearity and low voltage/current performance, it is available in a variety of surface mount packages suitable for a wide range of applications. Just as importantly, it can be manufactured cost effectively such that GPS based consumer electronic applications can certainly be brought within reach.

A two-stage, single-ended version of the design is presented, due to its applicability in the majority of GPS systems. This amplifier achieves greater than 26 dB of gain and less than 0.8 dB of noise figure at 1.575 GHz. It also exhibits excellent gain and noise figure characteristics from 800 MHz to 2000 MHz. The performance specifications for the amplifier are given in Table 1.

## GMIC Technology

GMIC circuits are produced utilizing a batch process similar to that utilized by the MMIC industry, with the exception that no active devices are produced on the circuit (1). All passive components such as resistors, capacitors, and induc-

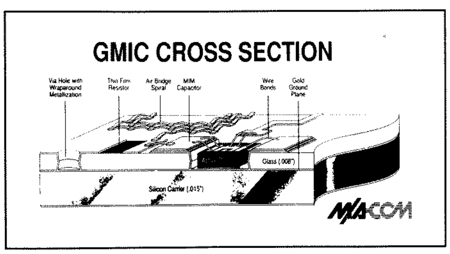


**Figure 1. Typical GMIC circuit cross-section with device mounted in via hole.**

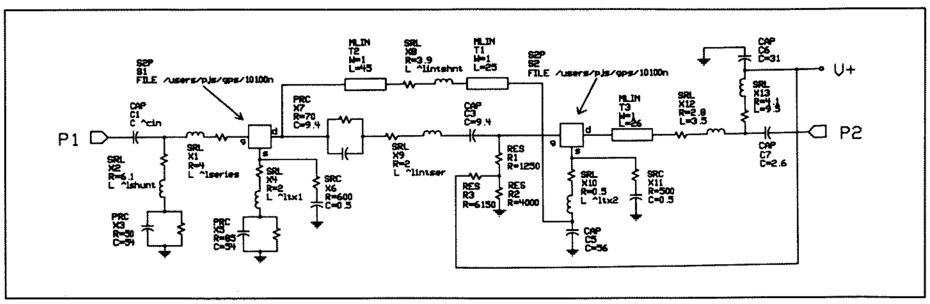


**Figure 2. Matching and bias circuitry of the GPS LNA.**

tors are defined and processed utilizing traditional multilayer thin film technology. The standard GMIC process consists of an 8 mil thick glass wafer laminated to a highly doped silicon wafer. The silicon wafer acts as a ground plane for the microwave signals as well as ground returns for DC.

The circuit is stepped and repeated on the wafer such that many circuits can be batch processed to achieve economies not realizable in typical hybrid circuits. The process also allows for air bridge interconnects which significantly reduce the need for wire bonds and allows for the integration of DC bias circuitry onboard the same substrate. The elimination of wire bonds increases the yields of interconnects, reduces RF circuit discontinuities, and improves the repeatability of RF performance. Wire bonds are still used for the integration of active devices into the GMIC. However, the number of wirebonds required in a GMIC design is significantly lower than in a typical hybrid design.

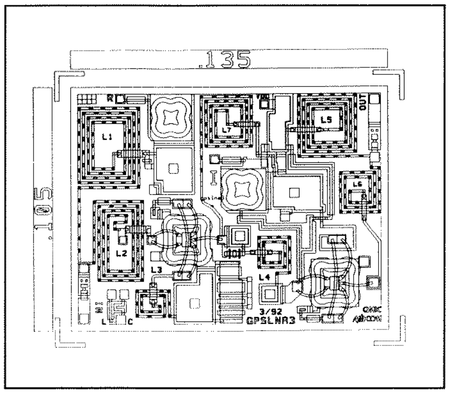


**Figure 3. GMIC circuit layout with FET assembly and wire bonding.**

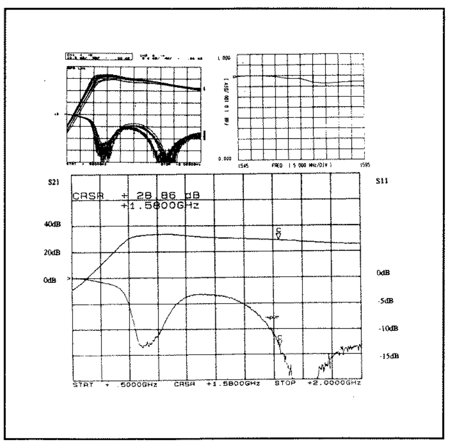


**Figure 4. Small signal performance of the 2-stage GPS LNA . The insets show the noise figure of the unit and the performance of 12 packaged amplifiers.**

The active devices are mounted within via holes which are etched into the glass dielectric and are attached to the underlying silicon wafer. The devices can be loaded into the GMIC using industry standard automated die placement machinery while at the wafer level. The process consists of dispensing epoxy into each via hole followed by the insertion of the die into the via holes. Special precautions must be observed during the placement of the die due to the close proximity of the walls of the via hole and the die edges. The assembled wafer is then wire bonded with an automated wedge bonder. Each assembled wafer is then diced and packaged for testing if required. The wafer can be designed with RF probe reticles in place at every circuit site or at designated probe reticles. Processing the circuits at the wafer level provides the advantage of processing many circuits simultaneously, gaining the advantages of economies of scale. However, there is a tradeoff between the number of RF probe reticles and the number of circuits which can be placed on a wafer due to the real estate taken up by the probe reticles. A typical example of a GMIC circuit with devices inserted into via holes is shown in Figure 1.

A single amplifier takes up 0.135 x 0.105 inches of GMIC real estate. This reticle is stepped and repeated across the useable area of a 3 inch wafer. This results in approximately 350 circuits per wafer, each of which is DC tested with an automated probe station and marked. The active devices are then inserted only into the circuits which pass this electrical test. The wafers are then diced, packaged, and a final test is performed before shipment. Before dicing, the circuits can be submitted for on-wafer RF testing if required.

## Design

The design goals for this amplifier were to achieve less than 1 dB of noise figure and a return loss of better than 10 dB. Also, a gain of greater than 25 dB was required with the amplifier being unconditionally stable. A single ended approach was taken due to the noise figure requirement. An Avantek MESFET device (ATF-10100) was chosen for cost and performance considerations, however the design has been proven with other FETs.

The design of GMIC circuits can be accomplished with both lumped and distributed elements, depending on the designer's requirements. For lumped element designs, an element layout library exists within customized CAD tool software in which the designer can select standard library elements to accomplish the design. S-parameter data is available for all the standard library elements. In many cases, where the designer requires elements not included in the library, approximate circuit models for non-standard elements are also available. These elements are useful up to the resonant frequency of the element and allow the designer to customize his layout for better real estate utilization.

The requirement of simultaneous gain and noise-match present conflicting design requirements in a single ended LNA design. Series feedback (2) in the form of a source inductor was added to the first stage to bring the optimum noise figure impedance closer to the ideal match impedance. The source inductance also has the effect of increasing stability, decreasing gain, and to some degree increasing the noise figure due to the series resistance of this inductance. Therefore, there are

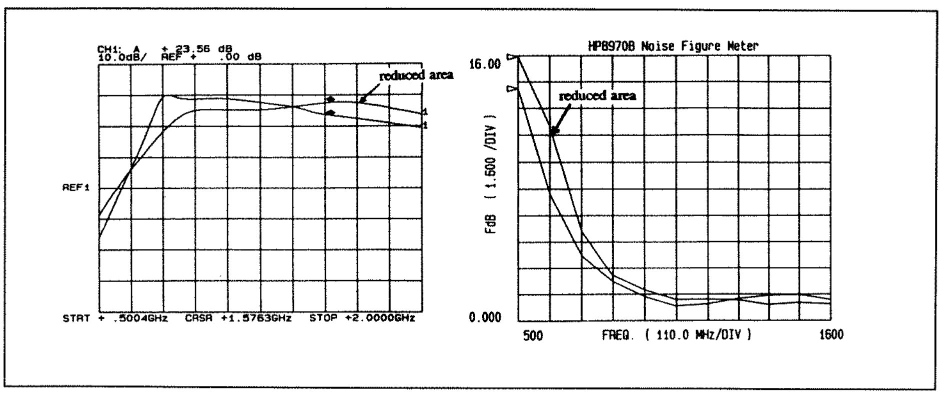


**Figure 5. Comparison of RF performance of the original 2-stage design and the new reduced area design which is 40 percent smaller in size.**

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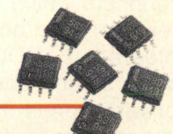

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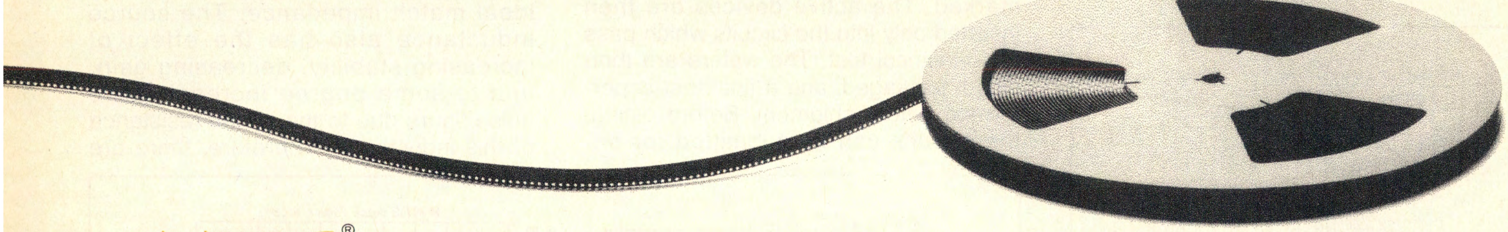

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tradeoffs which must be considered for this type of design. The noise figure requirements for this amplifier limit the variety of design approaches which can be utilized.

Due to the repeatable nature of the GMIC process, a lumped element matching approach was considered. This approach has an extreme size advantage over distributed elements at the frequency of interest (3). It also allows incorporation of the bias network into the matching structure. The matching structures were selected as shown in Figure 2 to achieve the desired matching for the circuit as well as provide DC and RF blocking elements for each stage. Resistive feedback was incorporated into the amplifier to ensure stability outside the frequency of operation without sacrificing gain and noise figure performance in the operating frequency range.

A self-bias approach was used so that only one supply voltage would be required. In the interest of minimizing power consumption, the FETs were biased in series and the two stages were decoupled with the use of bypass capacitors. The bias supply voltage range for the amplifier is 3.5 volts to 6.0 volts.

Considerable attention was paid to the layout due to the size restrictions imposed by the requirement that the amplifier fit into a small eight-lead surface mount package. The matching elements double as bias chokes to minimize the number of elements, and consequently, the size of the circuit. The number of interstage elements were reduced by matching the output of the first stage to the complex conjugate of the input of the second stage. The layout of the amplifier is shown in Figure 3.

## Performance

The circuit was modelled on a conventional linear analysis simulator. The results of this simulation and the measured data are given in Figure 4. The amplifier was tested both on wafer and in various packages and fixtures. The actual performance is very consistent with the simulated results. The amplifier is unconditionally stable and can be operated from a single voltage supply over a range of 3.5 to 6 volts. Across this voltage range, the maximum DC bias current is 25 mA. Noise figure over the range of –40 to +80 degrees C and at 1575 MHz is less than 1.3 dB. The gain over this temperature range is greater than 25 dB. No significant degradation was observed in various sealed glass and ceramic packages.

Note that the amplifier has a dual band performance at 800-900 MHz and 1400-1600 MHz with usable gain in the 700-2000 MHz range. Nominal noise figure at 1575 MHz is 1.0 dB and at 900 MHz is still 1.2 dB. A reduced area version has been fabricated and early results for the smaller GMIC design are shown in Figure 5.

Single stage versions of this circuit have also been designed and built with good success. These circuits have the same footprint as the two stage version and have gain and noise figures listed in Table 2.



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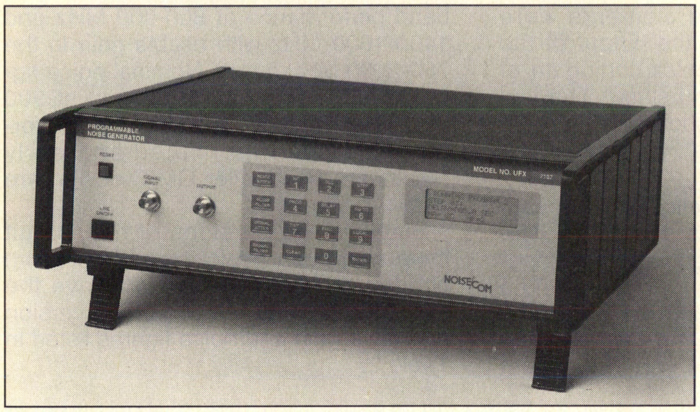


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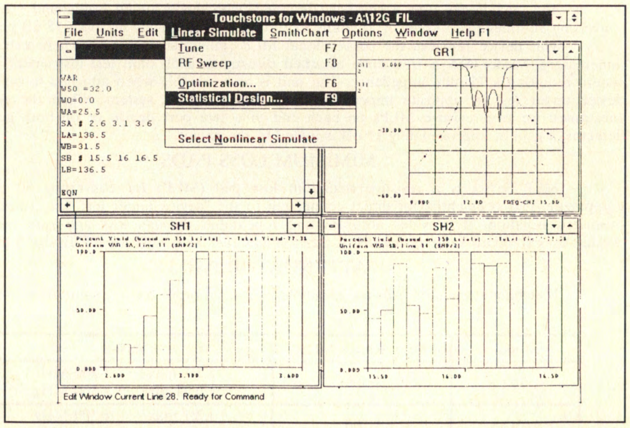


# Watch for more prizes in the next issue of *RF Design*!

| | LNA#1 Single Stage | LNA#2 Single Stage | LNA#3 Two Stage |
|---|---|---|---|
| Frequency | 1575MHz | 1575MHz | 1575MHz |
| Gain | 12dB | 14dB | 26dB |
| VSWR | 1.5:1 | 1.5:1 | 1.5:1 |
| Noise Figure | 0.8dB | 0.8dB | 0.8dB |
| Current | 25mA | 12mA | 25mA |
| DC Voltage | 3.0-6.0V | 3.0-6.0V | 3.5-6.0V |
| P1dB | +10dBm | +9dBm | +10dBm |
| IP3 | +20dBm | +18dBm | +20dBm |
| Size (L x W) | 0.250" x 0.190" | 0.250" x 0.190" | 0.135" x 0.115" |

**Table 2. GPS LNA performance**

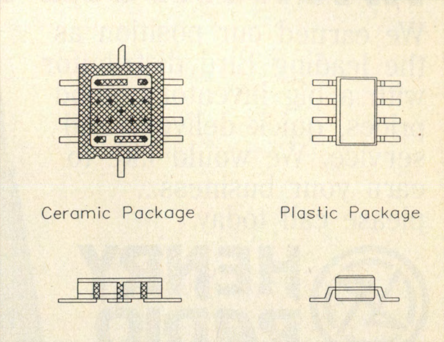


**Figure 6. Packaging of the GMIC amplifier.**

## Packaging

Since the thrust of this design is the commercial market, a low cost packaging alternative is required. Most component users desire a packaged amplifier for ease of handling and various other reasons, so several packaging options have been explored. A low cost surface mount ceramic packaging approach is currently being pursued while plastic packaging options are also being developed.

The ceramic package consists of a metalized ceramic base-ring fired together with a metal lead frame brazed to the bottom. The bottom ceramic has a metallized pattern screen printed on both the top and bottom surfaces. Top side metallization is used for circuit attachment while bottom side metallization is for lead frame attachment. The leads of the package are an iron-nickel alloy which are attached to the base with a copper silver braze. Overall package size is 0.315 x 0.350 inches including lead lengths. With the cover attached the thickness of the package is 0.075 inches.

A hermetic seal is achieved by attaching an iron-nickel alloy cover to the top of the seal ring. This cover has a gold tin preform which is tack welded to it for the sealing operation. The preform is then reflowed in a nitrogen atmosphere with the use of a dap sealer. A solid tungsten filled via provides a transition between the package leads and the inside of the package. Electrically, the lead structure is coplanar on the outside of the package and microstrip on the inside of the package. This structure is well understood and provides adequate VSWR and insertion loss up through the upper end of the frequency band as well as providing a hermetic transition into the package.

A plastic packaged surface mount version of the amplifier (Figure 6) is also currently in development and will soon become available. This version will incorporate the GMIC attached to a metal lead frame with plastic injection-molded around the device. The body size of this package is 0.155 x 0.195 inches with leads extending 0.045 inches on either side of the body of the package. The plastic package footprint is then 0.245 x 0.195 inches. The thickness of the package is 0.058 inches, and the leads are formed in a gull wing configuration to allow for surface mounting of the device. Special attention must be paid to the circuit design and fabrication to make it suitable for the transfer molding process. Some degradation in the noise figure characteristics is expected in this package.

## Conclusion

The GPS amplifier described here utilizes GMIC technology to gain considerable size and manufacturing advantage. The designer is able to tailor the performance of the amplifier to the application by choosing the best active device and designing the matching and bias structures around the device. This complete integration is then placed in a variety of packaging options which best suits the individual customers needs. The resulting amplifier is a cost effective device which can be utilized in commercial and military GPS systems or various other applications in this frequency band. ***RF***

## Acknowledgements

The authors would like to thank the many people who made technical contributions which had a direct impact on this work. Specifically, A. Douglas for his design contributions, R. Perko for his technical insights, S. Ravid for his influence in packaging, and W. Foley for the never ending support in the test area.

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### About the Authors

Paul J. Schwab received BSEE and MSEE degrees from Pennsylvania State University. For the past four years he has been with M/A-COM as a senior engineer and in the last two he has been involved in the development of GMIC technology. He can be reached at 110 Chelmsford St., Lowell, MA 01851, or by phone at (508) 453-3100 X4674.

Joan Williamson has a Mechnical Engineering Technology degree from the University of Massachusetts, Dartmouth, formerly S.E. Mass. University. She can be reached at 202C Royal Pines Dr., Huntsville, AL 35806, or by phone at (205) 895-9327.

# RF software

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Compact Software has announced Super-Spice™, an RF, microwave and high-speed frequency/time-domain simulator working in UNIX X-Windows. This schematic-driven simulator provides accurate models of transmission media and discontinuities including bends, steps and crosses as well as an "electromagnetics module" which allows multi-layer analysis for multicoupled structures.
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## Worksheets for Communication Systems

CommSys is a Microsoft Excel worksheet for satellite communications system design. CommSys calculates cascaded noise figure, 1-dB compression points, G/T and S/N ratios, link budgets and in-band spurs. It also performs phase noise integration and antenna and transmission line calculations. CommSys is $40.00 and requires version 2.X or greater of Microsoft Excel for the Apple Macintosh or IBM PC.
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## Alaskan Topography Data

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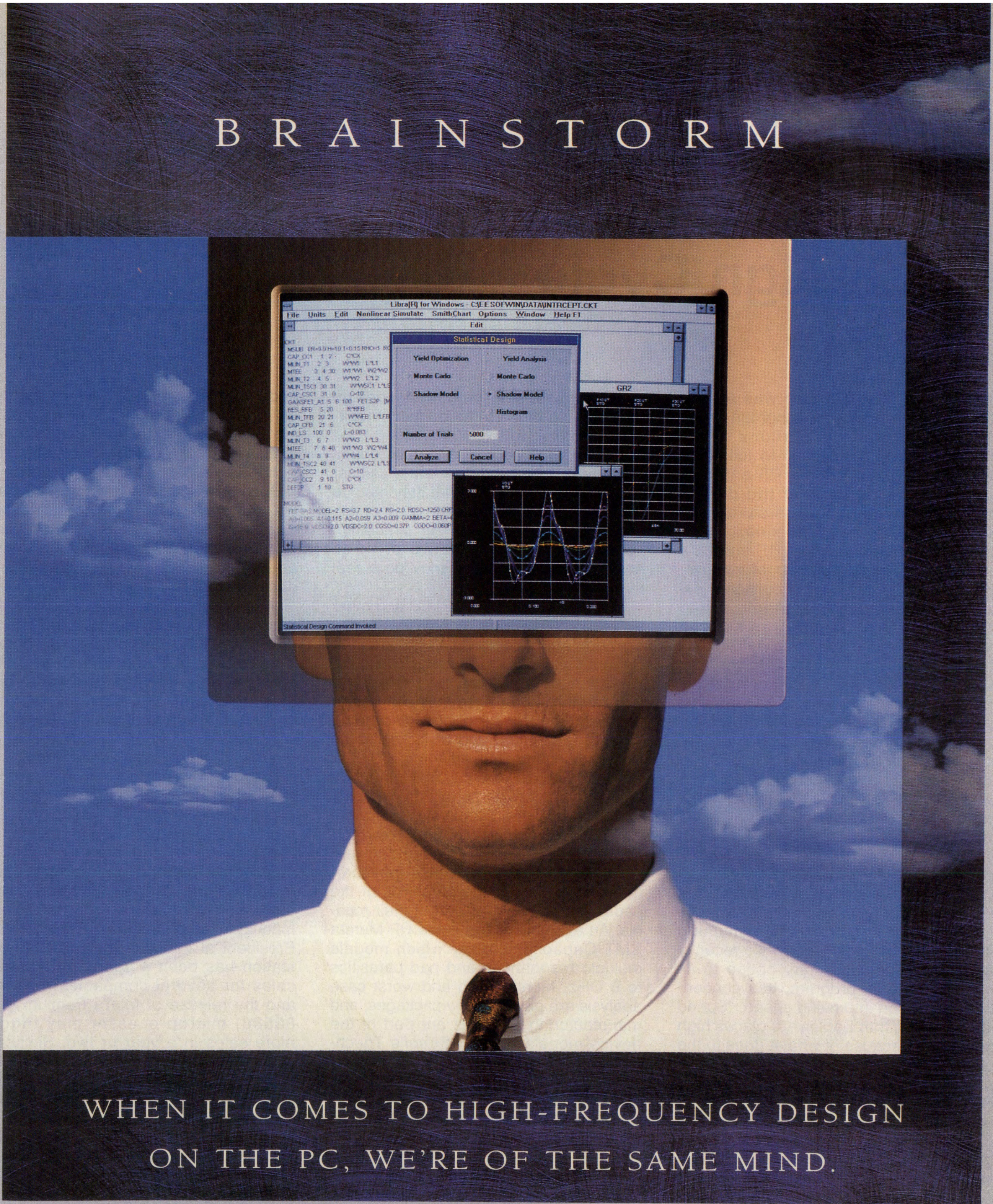

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# Software Revs Up the RF Design Cycle

*By Andy Kellett*
*Technical Editor*

In the past, RF software was a rope bridge crossing the treacherous, most calculation-intensive portions of the path from design inception to full scale production. Today, some RF software spans entire portions of the design cycle, while others provide sturdy and easy passage across smaller obstacles.

Many manufacturers of RF software are currently in the process of integrating schematic capture, circuit analysis and manufacturing design into single packages. This change was seen in recent versions of several well established software packages. PSpice, from MicroSim, was a stand-alone analysis tool just a few years ago. Now, as part of the Design Center family, it can be used in conjunction with schematic capture, waveform analysis, Monte Carlo analysis, PCB layout and other tools in a single seamless package. Similar functions are provided by Intusoft's ICAP/4, incorporating IsSpice; and Series IV are from EEsof, which incorporates Touchstone and/or Libra. The Hewlett-Packard RF Design System, first released this Fall, was born an integrated package.

Other manufacturers are remaining more focused. Eagleware, while recently adding a schematic entry option to its =SUPERSTAR= program, has concentrated more on improving speed. "Speed is something that Eagleware places high priority on because of what it does to the user's productivity," says Randy Rhea, president of Eagleware. *em* from Sonnet Software is a program devoted to calculating electrical parameters for 3-d planar conductors. "As a small company you have to be highly focused to be viable. That has been explicit since day one; we're focused exclusively on electromagnetic software," says Sonnet president James Rautio. According to Tesoft president Steve Lafferty, TESLA, his company's non-linear, time-domain, block-level simulator, "fills in a missing link between low-level, non-linear simulators like Spice and the frequency domain type techniques."

Where does this software fit into a designers's work? Accelerated design cycles are making cut-and-try design methods too slow for many manufacturers. According to EEsof Marketing V.P., Tom Reeder, "We work with hundreds of customers who build the latest cellular phones and they tell us that they have to put out a new model every six months. They are depending on RF simulation software to do one or two or three design iterations before they do a real manufacturing prototype."

In a similar way, MIC, MMIC and other circuit designers can reduce the money spent on shrinking circuits. According to Sonnet's Rautio, "A designer can compress the area of a circuit by a factor of four or more and then iterate the design on the computer as opposed to going through fabrication and test. That is an incredible competitive advantage."

Again with an eye towards manufacturing, RF software models are increasingly realistic, and analysis takes into account device statistics. Hewlett-Packard's RF Design System software libraries are based on actual measurements and include the parasitics of the pads and materials as part of their models. An example of this is the HP Murata SMT Capacitor Library which models multiple resonances and pad parasitics to 6 GHz. Monte Carlo and worst case analysis are part of many packages, and manufacturing statistics are part of the device models used in EEsof's Touchstone.

The limited scope of early software was not a result of a narrow vision of the usefulness of computers. In fact, according to the Marketing Manager for Hewlett-Packard's RF Design Software, John Hirsekorn, "We have had simulation tools inside H-P since the sixties, and good ones." The problem was that the computers required were enormously expensive, and even then, were less powerful than today's desktop computers.

The memory and speed available to engineering workstations, and even PCs, has made highly integrated and highly accurate software possible. Workstations use reduced instruction sets and high clock speeds to crank through calculation intensive applications like RF simulation. A personal computer employing a 486 processor overlaps the performance of low end workstations.

Software makers have not merely waited for the machines to get faster. Eagleware's =SUPERSTAR= v.4 employs two port analysis, model caching and other algorithmic methods to produce virtually real-time linear simulation. "There is something that happens in the interaction between the user and the circuit when the interaction is real time. You will find the user tuning parts by tapping the keyboard, and very soon they are saying, 'Oh, I see what's going on.' and that in and of itself may suggest what to change in the circuit to change the result." says Eagleware's Rhea.

The users of design software are still requesting more speed, higher accuracy, easier use and more synthesis capabilities. Vendors are trying to respond to these requests and are preparing for new hardware and operating system introductions. "It seems people buy more software when they get a new machine," says Charles Hymowitz, Vice President at Intusoft. The HP 700 work station has been a source of software sales for several companies recently, and the release of Intel's Pentium (aka 80586) microprocessor may trigger more sales in a year or two. Similarly, several software vendors are waiting for Windows NT, which will allow much improved multitasking and interprocess communication.

Like all tools, software should be matched to the job. However, the nature of RF software has changed from a troubleshooting and number crunching tool to something more general. While not everyone can afford, or even needs an integrated design package, everyone can find software that will bridge the chasm they need to cross. *RF*

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# RF design MARKETPLACE



## RF engineering opportunities



### Advertising Index

# *RF literature*

## Catalog
Trilithic has released its 1993 catalog covering filters, attenuators, test components, switches, and switching and control subsystems. Technical specifications and ordering information are included in the 93-page catalog.
**Trilithic**
**INFO/CARD #159**

## Filter Market Report
A report entitled, "Electronic Filters — U.S. Markets, Applications and Competitors: 1992-1997 Analysis", has been published by World Information Technologies. The report contains information regarding consumption, production, product type, competition, imports & exports, applications, market shares and more. One complete report costs $1595, with additional copies for $159.
**World Information Technologies, Inc.**
**INFO/CARD #158**

## Chirp Synthesizer
Sciteq's booklet, "DCP-1 Chirp Synthesizer For Linear FM (LFM) Waveforms", discusses the advantages and applications for linear FM modulation, figures of merit for LFM synthesizers, technical specifications of the DCP-1 and performance data. The booklet is 32 pages long.
**Sciteq Electronics, Inc.**
**INFO/CARD #157**

## Dielectric Resonator Oscillators
A brochure from EMF Systems describes noise in dielectric resonator oscillators (DROs) and coaxial resonator oscillators (CROs) and includes specifications for their Series 600 phase locked DROs and CROs and their Series 500 free running DROs.
**EMF Systems, Inc.**
**INFO/CARD #156**

## EIA Standards and Publications
Global Engineering Documents offers an expanded, 120-page catalog of all standards, specifications, and other publications of the Electronic Industries Association (EIA) and its affiliated groups and divisions. The catalog includes publications of the Telecommunications Industry Association (TIA), the Joint Electron Device Engineering Council (JEDEC), the EIA Tube Engineering Advisory Panel (TEPAC), and other EIA segments. The catalog is free of charge.
**Global Engineering Documents**
**INFO/CARD #155**

## ITU Documents On-Line
TELEDOC, an electronic document distribution service providing remote access to International Telecommunications Union (ITU) documents, is now operational. The database currently contains CCITT and CCIR administrative documents, substantive input and proposals to CCITT and CCIR study groups, lists of CCITT reports and recommendations, CCITT and CCIR meeting schedules and other information.
**ITU**
**INFO/CARD #154**

## Component Database Upgrades
D.A.T.A. Business Publishing has released the *1992 Winter Quarter High Reliability Electronics Components D.A.T.A. DIGEST* and has expanded the *Integrated Circuits Library* and *Discrete Semiconductors Library.* The annual subscription price for four quarterly issues of the DIGEST is $185.00. Prices for the various D.A.T.A. DIGEST Libraries range from $99 to $205.
**D.A.T.A. Business Publishing**
**INFO/CARD #153**

## Crystal Catalog
Tele Quarz has released a catalog presenting their lines of quartz crystals, oscillators and filters. Included in the 32-page catalog is a new line of SAW filters. Electrical specifications and a dimensional drawing are included for all listed products.
**Tele Quarz GmbH**
**INFO/CARD #152**

## SPICE Newsletter
The *Intusoft Newsletter* is a publication dedicated to discussing topics related to the SPICE analog circuit simulation program. Newsletter subscribers receive a newsletter and diskette six times a year. The diskette contains symbols, schematics and models discussed in the newsletter, as well as related circuit design applications. The *Newsletter* is free of charge to anyone, but only subscribers receive the diskette.
**Intusoft**
**INFO/CARD #151**

## Filter Catalog
Micro-Coax™ has released a 38-page catalog on its full line of in-cable filters. The catalog provides technical data, ordering and service information, as well as a guide to specifying and selecting filters. Among the filters detailed are lowpass, highpass, video transient suppression highpass, bandpass and broadband (octave) bandpass filters.
**Micro-Coax Components, Inc.**
**INFO/CARD #150**

## Synthesized Signal Generators
A 14-page brochure describing the MG3631A and MG3632A synthesized signal generators has been released by Anritsu Wiltron Sales Company. The brochure describes the functions, performance characteristics and specifications of the generators. Peripheral equipment information and ordering information are also included.
**Anritsu Wiltron Sales Co.**
**INFO/CARD #149**

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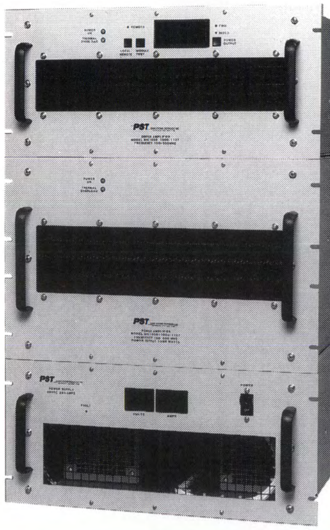

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| AR4819-10 | 400-1000 | 15 |
| AR4819-25 | | 40 |
| AR4819-50 | | 75 |
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