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July 1993

1993 *RF Design* Awards
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Terry Hock and Ray Page

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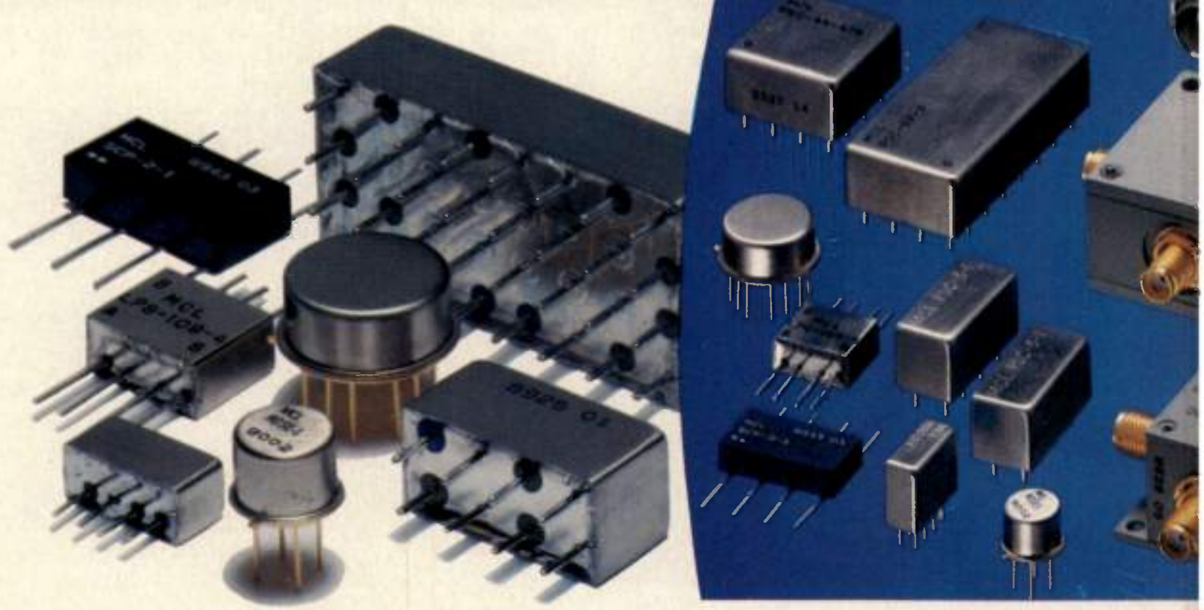

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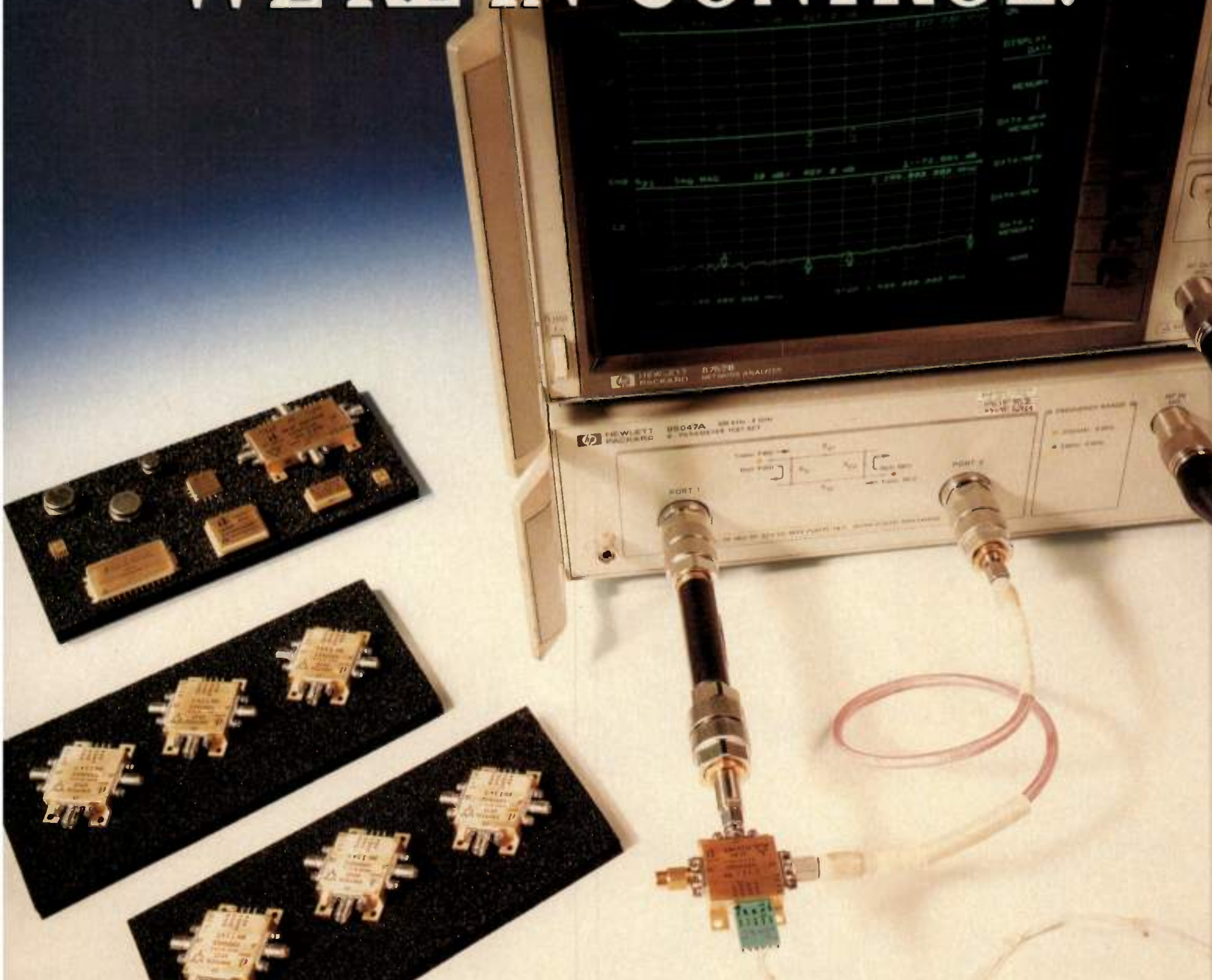

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SP2T	10-2000	0.60	82	1.000	TTL	14 Pin DIP	DSO052
SPST	10-2000	1.8	67	0.035	TTL	14 Pin SMP	DSO790
SP2T	DC-2000	0.7	50	0.200	TTL	TO-5	DSO813-T
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7	DC-1000	0.5-63.5	7.1	0.100	TTL	38 Pin DIP	DAO897
7	10-1000	0.5-63.5	6.4	0.035	TTL	SMA	CDAO867

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RF design

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July 1993

design awards

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This circuit, the design Grand Prize Winner of the 1993 RF Design Awards Contest, is a passive transponder intended for automated railcar tracking. Illuminating this low parts-count device with low power 915 MHz energy yields a modulated return signal at 1830 MHz.

— Raymond Page

37 Synthesizer Design With Detailed Noise Analysis

The winning software entry in the 1993 RF Design Awards Contest will help analyze and design phase locked loop synthesizers. The program provides a thorough model for all contributions to phase noise within the loop and predicts spur levels. Changes to loop response parameters result in an immediate change in the on-screen response graphs.

— Terrence F. Hock

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49 Contest Winners Demonstrate Innovation and Hard Work

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64 RF Component Modeling for CAD/CAE

This tutorial presents lumped passive models for RLC components, including parasitic effects and losses. For simplicity, only first order resonances are considered, although second order effects are discussed. *— Les Besser*

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Put the "R" Back Into R&D

By Gary A. Breed
Editor

Here in the U.S. we have a remarkable history of making scientific discoveries and engineering breakthroughs. We take pride in putting men on the moon, developing super computers, and creating world-wide communications systems.

But many American innovations have become commercial successes for foreign companies. Business practices in other countries often emphasize manufacturing rather than research. In the name of competitiveness, we're seeing U.S. industry being pushed to do the same; putting all of its effort into development. We may be over-reacting, with the danger of short-changing research.

The computer industry is giving us a look at the cost of a one-sided approach. Many companies that were riding on the PC boom are now in trouble as growth slows in the one technology they have been exploiting. Even big computer companies seem uncertain about the next step they should take. Maybe they don't have enough new ideas fermenting in their labs.

It's not time to panic, but the pendulum needs to start swinging back toward center. Fortunately, the timing of new RF growth opportunities puts them in the development stage and the current attitude is not yet a problem. But before we get caught unprepared for the next generation of opportunities, let's be sure that new ideas are being explored.

Some of these new ideas have been explored by engineers seeking recognition in this year's RF Design Awards contest. Ray Page's winning RFID transponder scheme is a good example of engineering research and development, while Terry Hock's championship PLL design and analysis program demonstrates how engineers create tools to support their R&D work.

Our contest has been around for eight years now, and I took some time to reflect on past winners. At least 75 percent of the time, the winning engineers had significant, direct support from their employers. This support ranged from giving complete freedom to develop the project, to providing access to test equipment and time for writing. Yet, of the 11 Grand Prize winning entries, only three were the result of regular job assignments. There's a message here that shouldn't be missed — creative ideas are developed in an environment that supports creativity. If those companies had limited our past winners to only "productive" work, we would not have seen their great ideas.

To sum things up, it is essential that the U.S. electronics industry learn better ways to bring products to market. However, this should be a *new* capability, not a replacement for our proven record of innovation and creative research. We should be careful to encourage our best people to be creative, to maintain a constant and generous flow of new concepts and discoveries to fuel those newly-developed manufacturing skills.

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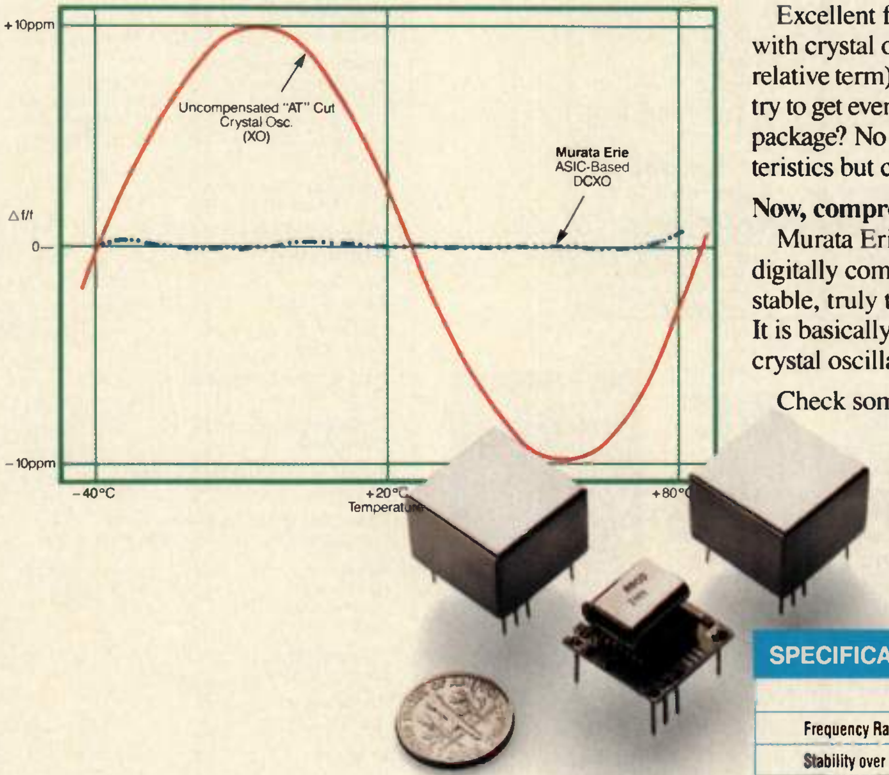

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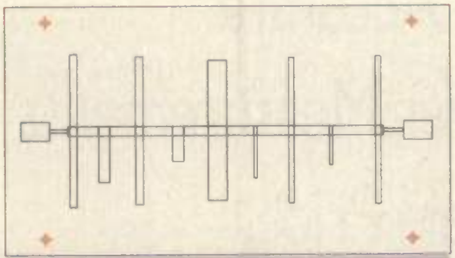
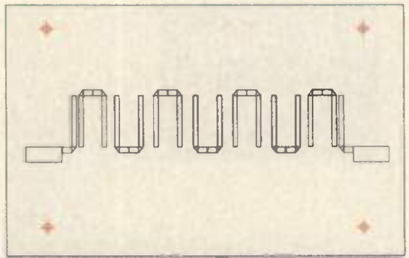
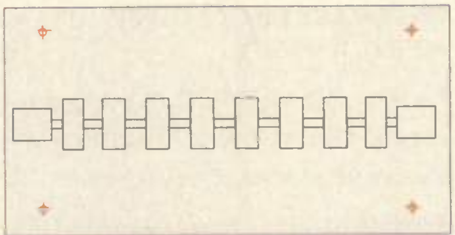
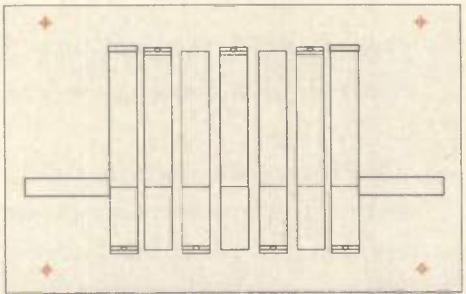
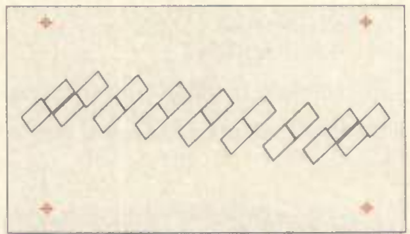

5.6 GHz Elliptic Bandpass on Teflon Board

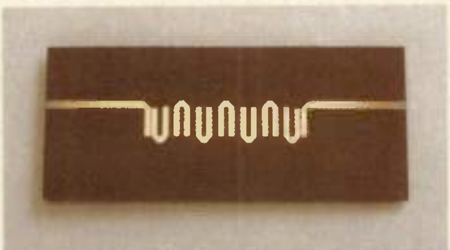

1.27 GHz Machined Combline Bandpass

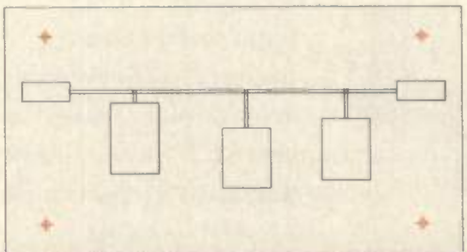

2.2 GHz Hairpin Bandpass on Teflon

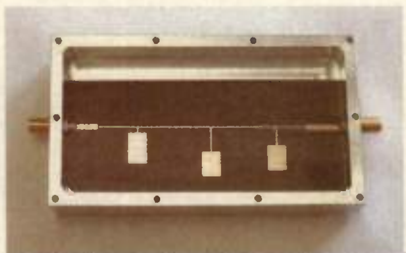

6.4 GHz High Order Stepped-Z Lowpass


10 GHz Wideband Interdigital BP


5.6 GHz Edge Coupled BP on Teflon


5.6 GHz Hairpin Bandpass on Teflon


1.2 GHz Elliptic Lowpass on G-10/FR-4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start to art

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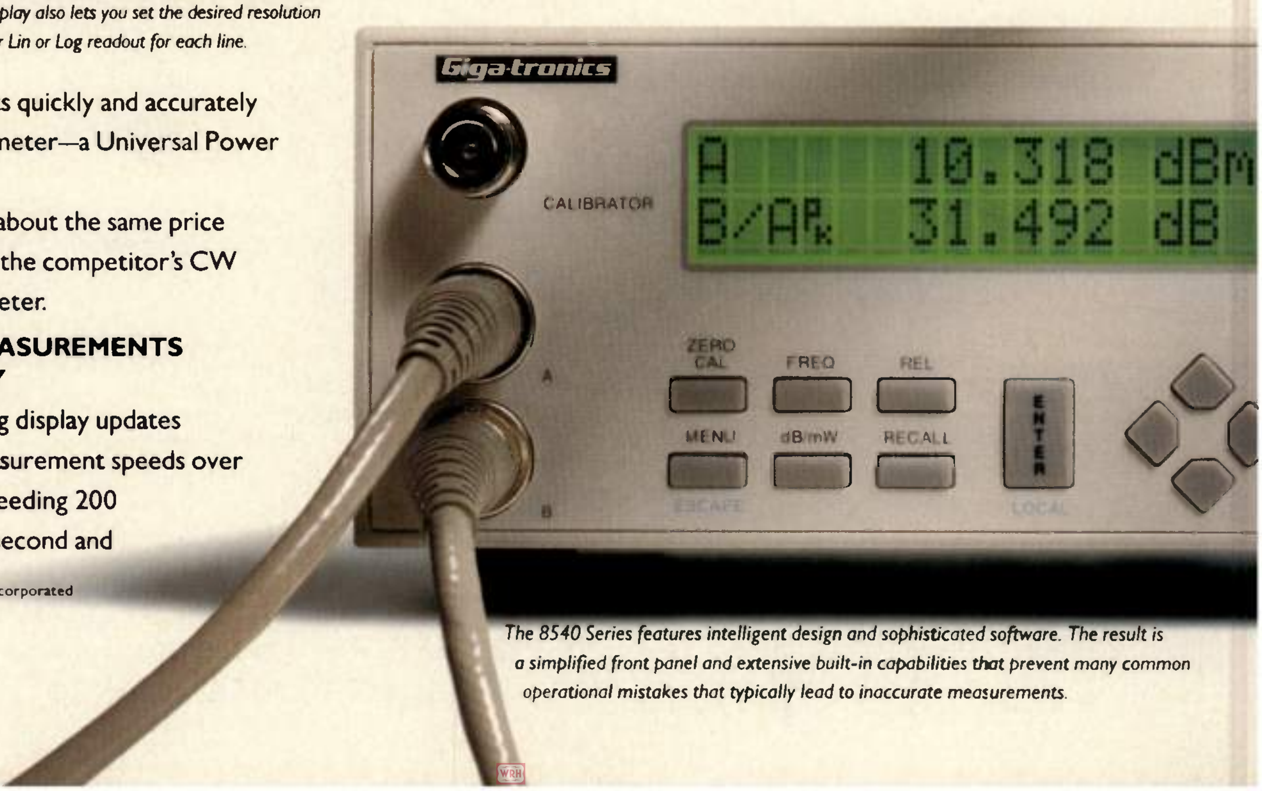
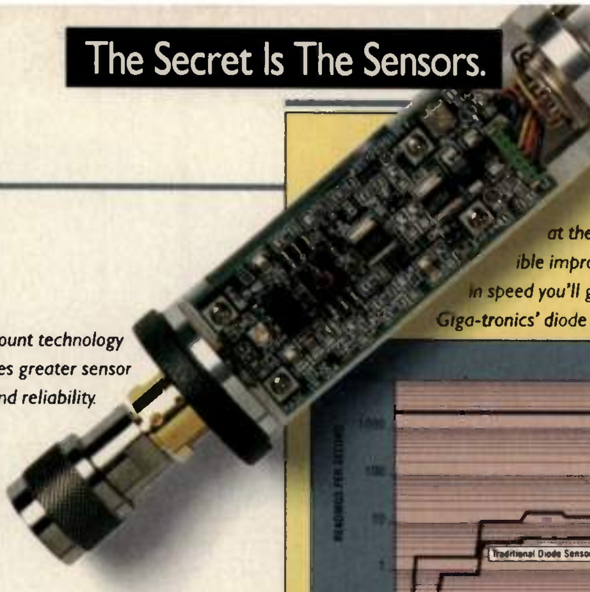


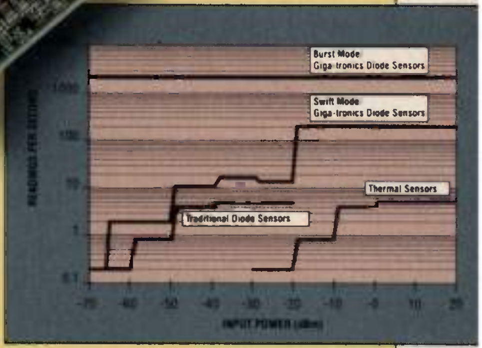
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

The Secret Is The Sensors.



Surface Mount technology assures greater sensor accuracy and reliability.

Just look at the incredible improvement in speed you'll get with Giga-tronics' diode sensors.



You'll be confident of your peak power readings, and still have all the benefits of an incredibly fast CW power meter.

ONE OR TRUE TWO CHANNEL OPERATION

If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, the Model 8542 lets you see readings from both channels *simultaneously.*

SIMPLE, INTELLIGENT OPERATION

The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

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The Giga-tronics 8540 Series delivers incredible performance by taking full advantage of the speed and dynamic range of diode sensors.

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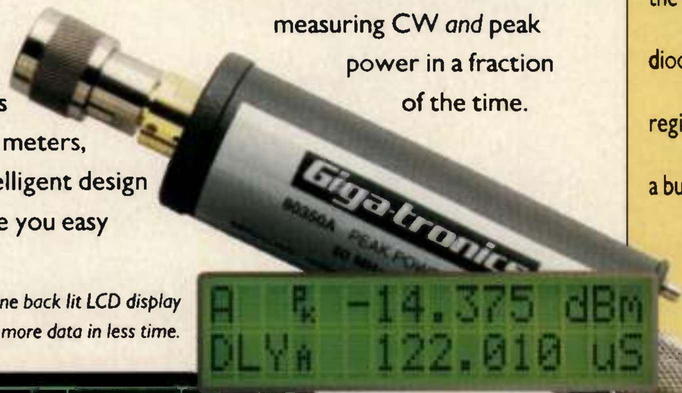
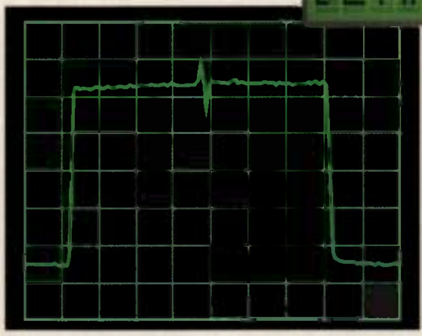


A two-line back lit LCD display provides you more data in less time.



The peak sensor adds a marker on a monitor output for setting an exact measurement point on pulsed signals.

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RF letters

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Proposed Signalling Technique Comes Under Fire

Editor:

The system described by Greg Magin ("A Robust Signalling Technique for Part 15 Control Network Applications", *RF Design*, April 1993) has virtually zero processing gain because the spread signal is amplitude limited prior to the despreading operation. In order to gain most of the potential benefits from spread spectrum, it is necessary that the received signal be despread and passed through a filter having the desired despread bandwidth PRIOR to any limiting operations. ...

John MacConnell
Itron/Enscan
Eden Prairie, MN

Editor:

... Assume for a moment that the jammer power level was 1 dB above the power level of the desired signal. Since the system is hard limited, the output of the limiter would be the squared up jammer containing none of the information from the desired signal other than zero crossing timing jitter. ... The important point to understand here is that once the jammer signal level is increased by 1 dB over the desired signal level, the hard limiting of the system is going to erode ALL of the process gain. ...As Mr. Magin points out, the FCC is interested in the ability of the receiver to accept (put up with) and reject interference. The described architecture cannot reject any interference if it is stronger than the signal of interest.

John Kesterson
Mitsubishi Electronics America
Nevada City, CA

Mr. Magin responds:

... Most of the concern seems to focus on the use of post limiting despreading and signal processing. The architecture described in the article is but one possible implementation of the signalling technique. ... The post limiting architecture was chosen for low cost receivers because of the many real advantages including: reduced component count and cost, reduced power consumption and elimination of AGC issues.

The major concern seems to be that when the jammer power equals the signal power, the limiter is captured and all information is lost. Our system produces very acceptable performance when the jammer power is equal to the signal power. Complete message packets are typically received with an 80 percent success rate. This corresponds to a BER of 3.4×10^{-3} which is more than 10 dB better than that achieved by a coherent BPSK system under similar conditions. It is interesting to note that the limiter-based Intellon system performed on par with classical direct sequence pre-detection despreading systems during jammer testing performed by the EIA CEBus RF committee. Pre-detection despreading systems must achieve and maintain accurate carrier lock in the presence of interference. This typically adds a lot of overhead in terms of synchronization preamble time and proves to be the weak link in many implementations.

To prevent loss of information when the level of the jammer exceeds the level of the desired signal, the CEBus RF system transmits the desired information on two identical sidebands which are separated by 10.5 MHz. When CRC errors are detected on one sideband, the receiver automatically tries the other sideband. For this reason, an interferer may be many orders of magnitude stronger than the desired signal without affecting link performance. The merit of a low cost communication link should be based on overall system performance and cost. We believe that our system achieves an extremely high performance to cost ratio.

Greg Magin
Intellon Corp.
Ocala, FL

Correction

In Andrzej Przedpelski's May 1993 tutorial,"A Comparison of Simple Software Methods for RF Calculations", equation three should be:

$$[(42+j2\pi f \bullet 0.01+(j2\pi f \bullet 0.342\times10^{-15})^{-1} + j2\pi f \bullet 2\times10^{-12}]^{-1} \qquad (3)$$

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RF calendar

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9-13 **IEEE International Symposium on Electromagnetic Compatibility**
Dallas, TX
Information: Dr. Frederich M. Tesche, c/o International Compliance Corporation, 1911 E. Jeter Rd., Argyle, TX. Tel: (817) 491-3696. Fax: (817) 491-3699.

17-20 **Sixth Workshop on Technical and Scientific Aspects of MST/ST Radar**
Chung-Li, Taiwan
Information: Professor Shin-Yi Su, Chairman, Local Committee, NSC Chung-Li Instrument Center, National Central University, Chung-Li, Taiwan, R.O.C. Tel: (886) 3-4227151 ext 5750. Fax: (886) 3-4255535 and (886) 3-4224394.

18-21 **IEEE 1993 International Geoscience and Remote Sensing Symposium (IGARSS '93)**
Tokyo, Japan
Information: Mr. Natsuhiko Motomura, Remote Sensing Technology Center of Japan, 7-15-17 Roppongi, Minato-ku, Tokyo 106, Japan. Tel: +81-3-3403-1761. Fax: +81-3-3403-17 66.

18-21 **International Symposium of Radio Propagation (ISRP '93)**
Beijing, P.R. China
Information: Prof. Zong Sha, China Research Institute of Radiowave Propagation, P.O. Box 138/93, Xinxiang, Henan 453003, China.

24-26 **Advanced Microelectronic Qualification/Reliability Workshop**
Denver, CO
Information: Dorthy Kelly, GTS, Inc., 3100 Route 138, Wall, NJ 07719. Tel: (908) 544-3231. Fax: (908) 389-3992.

25-2 **XXIV General Assembly of the International Union of Radio Science (URSI)**
Kyoto, Japan
Information: Prof I. Kimura c/o Osaka Office, Business Center for Academic Societies of Japan, 10th Tabuchi Bldg, 6-3 Matsugaecho, Kita-ku, Osaka 530, Japan. Tel: (81) 6-356-6041. Fax: (81) 6-356-6190.

30-3 **European Conference on Circuit Theory and Design**
Davos, Switzerland
Information: Prof. J. Neirynck or Mrs. Renate Agotai. Tel: (41) 1-256-51-93. Fax: (41) 1-262-08-23.

September

6-9 **23rd European Microwave Conference and Exhibition**
Madrid, Spain
Information: Microwave Exhibitions and Publishers, 90 Calverley Road, Tunbridge Wells, Kent TN1 1BR, England.

RF courses

Basic Radar
July 13-15, 1993, Atlanta, GA
Near-Field Antenna Measurements and Microwave Holography
July 19-22, 1993, Atlanta, GA
Radar Design Workshop
August 17-19, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

High-Speed Communication Networks
August 16-18, 1993, Santa Cruz, CA
Compression Technologies: Image, Video & Associated Standards for Computers, Communications & Consumers
August 19-20, 1993, Santa Cruz, CA
Information: University of California Extension, Santa Cruz, Tel: (408) 427- 6600. Fax: (408) 427-6608.

RF Circuit Components: Measurements, Models and Data Extraction
July 12-16, 1993, Oxford, UK
August 16-20, 1993, Los Angeles, CA
RF Circuit Design: Passive and Active Linear Networks
August 24-27, 1993, Los Altos, CA
RF Design: Nonlinear Circuits and Devices
August 30-Sept 2, 1993, Los Altos, CA
Information: Besser Associates, Tel: (415) 949-3300. Fax: (415) 949-4400.

Finite Element and Finite Difference Time Domain Methods for Solving Electromagnetic Engineering Problems
July 19-21, 1993, Worchester, MA
Electromagnetic Compatibility and Interference
September 14-17. 1993, San Diego, CA
Information: Southeastern Center for Electrical Engineering Education, Kelly Brown - Registrar. Tel: (407) 892-6146. Fax: (407) 957-4535.

Inherently Conductive Polymers: An Emerging Technology
September 8-10, 1993, Boston, MA
Information: Advanced Polymer Courses, Dr. M. Aldissi. Tel: (802) 655-2121. Fax: (802) 655- 2025.

Wavelet Transform: Techniques and Applications
August 9-11, 1993, Los Angeles, CA
Data, Speech, Image and Video Compression: Principles, Applications and Standards
August 9-13, 1993, Los Angeles, CA
Active and Passive RF Components: Measurements, Models, and Data Extraction
August 16-20, 1993, Los Angeles, CA
RF and High Speed Digital Circuit Components: Measurements, Models and Data Extraction
August 16-20, 1993, Los Angeles, CA
Wireless Personal Communications Services
September 15-17, 1993, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825- 1047. FAX: (310) 206-2815.

INFO/CARD 125

Bit-Error Rate Tests With Variable E_b/N_o

High data rates and high reliability are of utmost importance in modern digital communication systems. BER versus E_b/N_o or BER versus $(C_o + N_o)/N_o$ is used as a quantitative quality factor when evaluating products, testing data links to determine if they will sustain higher data rates, and in verifying performance of digital communication systems on-site or in production.

BER versus E_b/N_o curves can be generated easily with a precision noise generating instrument and a BER tester that is often built into digital communication equipment.

The E_b/N_o of an RF, microwave, or fiber-optic digital communication system can be adjusted either before or after downconversion to an intermediate frequency. The procedures using Noise Com's UFX-BER Series noise test sets are outlined below.

The total power level is adjusted to the level specified by the receiver manufacturer and maintained at this level with a 1 dB (or 0.1 dB) step attenuator. The signal carrier power can be kept constant by passing the signal through a linear automatic gain controlled (AGC) amplifier. The amplifier must be fast enough to compensate for the fading encountered in RF and microwave links. The carrier power level is then set with the instrument's built-in precision step attenuator.

A precise calibrated amount of White Gaussian noise is injected at an IF, RF, or microwave frequency in the receiver or communications channel. The injected noise power level is adjusted only when necessary to maintain constant total power.

The instrument's built-in power meter assists in setting the carrier and the total power levels as well as performing a test of the noise level and attenuator settings.

The signal plus injected noise is now fed into the receiver's RF or IF input, and the BER is measured at 1different settings of E_b/N_o with the total power level kept constant. The E_b/N_o is a function of the bit rate, carrier power, and noise level as shown in the following equation:

$$E_b/N_o(\text{in dB}) = C(\text{in dBm}) - N_o(\text{in dBm/Hz}) - 10\log(f_b)$$

where:

f_b is the transmission bit rate
N_o is the noise spectral density
E_b is the bit energy
C is the carrier power

A correction factor for the noise level must be applied when the added noise level approaches the noise floor of the receiver:

$$E_b/N_o(\text{in dB}) = C(\text{in dBm}) - 10\log(N_i) - 10(f_b) - 10\log(1 + F \times k \times T/N_i)$$

where:

N_i is the injected noise spectral density generated by the noise test set (in mW/Hz)
T is the room temperature (in degrees K)
F is the noise factor at the injection point of the receiver
k is Boltzmann's constant $= 1.380 \times 10^{-23}$ J/K

If this correction is not added, the BER will be larger, a worst case value, than the result of a corrected measurement.

$(C_o + N_o)/N_o$ is related to E_b/N_o as follows:

$$(C_o + N_o)/N_o(\text{in dB}) = 10\log[(10^{[(E_b/N_o + 3)/10]}) + 1]$$

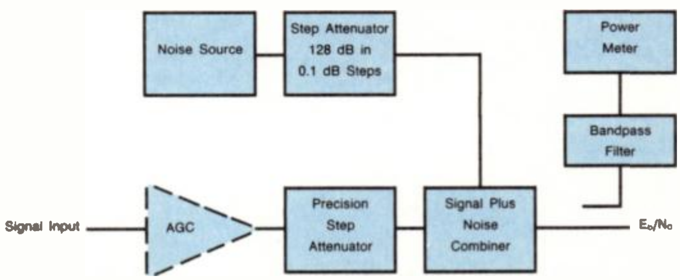


Figure 1. IF Noise Test Set for BER Vs E_b/N_o Measurements.

Noise Com's UFX-BER Series will automatically make this correction and display the actual E_b/N_o or $(C_o + N_o)/N_o$ once the transmission bit rate, f_b, room temperature (T), and receiver noise figure, NF = 10log (F), have been entered.

Turn the page for related products . . .

NOISE/COM
E. 49 Midland Avenue, Paramus, New Jersey 07652
(201) 261-8797 • FAX (201) 261-8339

E_b/N_o Noise-generating Instruments for Bit Error Rate Testing

UFX-BER Series

General Specifications

Carrier Path

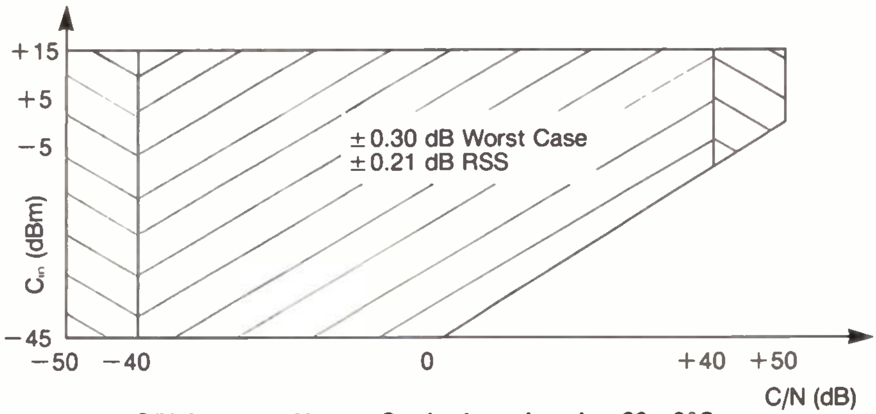
Attenuation	0 to 60 dB, automatically programmed
Range of carrier input power (C_{in})	−45 to +15 dBm
Flatness	±0 20 dB for 70 MHz (±20 MHz) ±0 30 dB for 140 MHz (±40 MHz)
Group delay	0 20 ns or less

Output Path

Range of carrier output power	−45 dBm to carrier input level
Range of noise output power (N)	−45 dBm to +5 dBm
Noise flatness	±0 50 dB or less from 100 kHz to 200 MHz
Accuracy of C/N	See graph
Third-order intercept point	+29 dBm typical
Crest factor (peak to average noise voltage)	14 dB or more

UFX-BER SERIES

MODEL	CENTER FREQUENCY	BANDWIDTHS
UFX-BER-455	455 kHz	Filters are customer specified Maximum bandwidth 200 MHz
UFX-BER-10	10.7 MHz	
UFX-BER-21	21.4 MHz	
UFX-BER-45	45 MHz	
UFX-BER-70	70 MHz	
UFX-BER-140	140 MHz	
UFX-BER-IF1	70 MHz, 140 MHz	
UFX-BER-__	User Specified	



C/N Accuracy Versus Carrier Input Level at 23±3°C

Features:

- Direct display of E_b/N_o or C/N
- Accuracy of 0.21 dB RSS
- 75 ohm impedance BNC connectors
- Variable output power

The UFX-BER Series of noise-generating instruments display the actual E_b/N_o or C/N for testing of satellite communications or digital radio equipment. The UFX-BER generates and measures E_b/N_o for IF back-to-back or loop-back testing with extreme precision over a broad range of input or output power, steps, increments, decrements, and desired E_b/N_o values can be entered directly in dB in 0.25 dB increments from the front panel or by IEEE-488 bus.

Settings are stored in non-volatile memory and the user can create up to eight test routines (including delay time and loops) that will run automatically under program control. The instrument measures 5 1/4 in. high, 17 in. wide, and 20 in. deep and comes with brackets for 19-in. rack mounting.

Options:

1. Automatic gain control amplifier to maintain constant carrier power level
2. Other custom configurations
3. 50-ohm input and output impedance instead of standard 75-ohm impedance
4. RS-232C, RS-422, or RS-423 interface in addition to IEEE-488 interface
5. 220 VAC, 50 Hz, instead of standard 110 VAC, 60 Hz
6. Interferor input
7. Switched filter bank with up to six customer-specified filters

NOISE/COM E. 49 Midland Avenue, Paramus, New Jersey 07652
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Canada Establishes Digital Radio

The Canadian Broadcasting Corporation (CBC) and the Canadian Association of Broadcasters (CAB) recently announced the formation of a new corporation, Digital Radio Research Inc. (DRRI) which will launch experimental digital radio service in Canada by late 1993. The DRRI will build and operate a number of transmitting installations to be used for digital radio research and development, and commission and publish research studies on digital radio broadcasting technology. The initial sites of the prototype stations will be in Montreal and Toronto. They will operate on the 1.5 GHz band (L-band) and begin broadcasting before the end of the year by using new or existing programming from the CBC and private broadcasters and converting it to digital form.

U.S. and Japanese Engineer Exchange Program — The U.S. Commerce department has announced a program to place U.S. engineers in Japanese manufacturing firms for up to one year. The goal of the Manufacturing Technology Fellowship project is to help U.S. engineers learn more about — and then use — Japanese manufacturing practices and to promote long-term professional exchanges with the Japanese. Fellowships will last about 15 months, including three months of intensive Japanese language and culture training in the United States. That will be followed by up to 12 months of hands-on manufacturing experience in Japan. U.S. participants will learn about Kanban, Just-In-Time manufacturing, Total Quality Control and other Japanese-inspired techniques as they are used in Japan.

Candidates must be U.S. citizens or permanent residents currently employed in manufacturing at an American company. They must have at least two years of manufacturing experience and long-term goals in manufacturing. For more information about the program or to receive an application, potential candidates should contact: The U.S. Japan Manufacturing Technology Fellowship Program, c/o Japan Technology Program, Room 4817, U.S. Department of Commerce, 14th St. and Constitution Ave., N.W., Washington, DC 20230. Tel: (202) 482-0356. Fax requests are encouraged at (202) 482-4826. Applications are due no later than July 16, 1993.

NIST Publication Available — A new NIST publication, *Measurements for Competitiveness in Electronics* (NISTIR 4583), identifies those currently unmet measurement needs most critical for the U.S. electronics industry to compete successfully worldwide. Nine fields of electronics are covered: semiconductors, magnetics, superconductors, microwaves, lasers, optical-fiber communications, optical-fiber sensors, video and electromagnetic compatibility. Each field's section contains a technology review, an overview of economic importance to the world market, a look at U.S. industry goals for competing internationally and a discussion of measurements needed to meet those goals. NISTIR 4583 is available for $52 (print) and $19.50 (microfiche) prepaid from the

National Technical Information Service, Springfield, VA 22161. Tel: (703) 487-4650. Order by PB 93-160588.

Russian Company Enters Western Market — Svetlana Electron Devices Manufacturing Corporation has announced that they will market their power grid and modulator electron tubes in the United States and other Western countries. Svetlana power grid tubes will be sold through Svetlana Electron Devices, Inc. Svetlana is Russia's largest power grid and modulator tube manufacturer. They will introduce tube types widely used in Russia but new to Western equipment designers and will also provide exact replacement, plug-compatible Western tube types. For more information contact Svetlana Electron Devices, Inc., 3000 Alpine Road, Portola Valley, CA 94028. Tel: (415) 233-0429.

Potential Standard for Wireless LANs — National Semiconductor recently proposed an open standard for Wireless LANs to foster interoperability between equipment offered by a wide variety of vendors. There are currently several Media Access Control (MAC) submissions before the IEEE 802.11 wireless LAN committee. One leading submission is a "Reservation-Based" MAC protocol based on a repetitive constant length frame which can support "isochronous" LAN traffic for voice and video. Another submission is the "Hybrid" Carrier Sense Multiple Access with collision Avoidance (CSMA/CA) protocol. The Unified MAC submission from National Semiconductor is based on major elements of both protocols. They are proposing a simple physical layer interface (PHY) description that partitions the synchronization, timing, and control functions on the MAC side of the service interface. This provides for a simple command set for the control of the radio side of the PHY to drive down both cost and power consumption.

Call For Papers — The Technical Program Committee of the Third International Conference on Multichip Modules, scheduled for March 29-31, 1994 in Denver, Colorado has issued a call for papers. The conference and exhibition is organized and sponsored by the International Electronics Packaging Society and the International Society of Hybrid Microelectronics. Topics being sought include: international and domestic applications, design and test, assembly and interconnections, materials, yield and reliability, processing MCM design and implementation, MCM/PCB interface, structure type, management and cost strategies, memory management, electrical and thermal performance, changes and change technology, telephone and cellular applications. Six copies of a 150 word abstract should be sent by November 1, 1993 to ISHM, 1861 Wiehle Ave., Suite 260, Reston, VA 22090.

"Clipper Chip" Developed for Communications Security — The National Security Agency with NIST assistance has developed the state-of-the-art "Clipper Chip" which can be used in new, relatively inexpensive encryption devices that can be attached to telephones. The microcircuit scrambles communications using a power encryp-

If you think the best RF Power Modules come from Japan, we'd like to re-orient you.

A lot of good things come from the "land of the rising sun"—like VCRs, hi-fi audio equipment and sushi. But when it comes to RF devices, specifically small amplifier assemblies called "modules" for use in cellular radios, the best source is right here in the good'ol USA. You got your first power modules for analog cellular radio from Motorola—now get your first power modules for digital cellular radio from the same source!

Motorola is proud to announce the first linear power modules, **MHW926** and **MHW927A/B**, for use in the United States Digital Cellular (USDC) radio system. Each module supplies 6 watts of average power, operates from 12.5 volts, and has the standard features of 50 ohm input and output impedance, stability, ruggedness, low harmonics and wide dynamic range in output control. The MHW927A/B require only 1 mW of RF input power, while the MHW926 requires 65 mW. Linearity is indicated by a 3rd order intermodulation distortion (IM3) specification of -29 dBc. A bias voltage of 9.5 volts is required for both the MHW926 and MHW927A, while the MHW927B requires a bias voltage of 8.0 volts.

Pioneers Then and Now!

Motorola was the first company to offer modules for mobile and cellular radios, having introduced modules for the 900 MHz cellular systems in the USA (AMPS) and in Europe (NMT) in the early 1980s. Last year we announced a complete line of linear modules for the digital cellular system in Europe (GSM). With the introduction of these linear UHF silicon bipolar modules, along with our recently introduced UHF GaAs FET power modules, we're proud to say we're continuing to take giant steps in cellular technology.

Check Out These Features:

- *Frequency— 824-849 MHz*
- *Output Power—6 W Avg.*
- *Input/Output Impedances of 50 ohms*
- *Supply Voltage—12.5 Vdc*
- *Bias Voltage—9.5 Vdc—MHW926 & MHW927A*
  8.0 Vdc—MHW927B
- *Input Power—18 dBm Max. for MHW926*
  0 dBm Max. for MHW927A/B
- *Harmonics— -30 dBc Max for 2Fo*
- *Package—301AB-01 for MHW926*
  301AA-01 for MHW927A/B

Whether you need tens, hundreds or thousands of modules, Motorola offers you repeatable performance from unit to unit. Plus quality, delivery and service you can depend on. So why go all the way to Japan for the Power Modules you need? Reorient yourself to an American pioneer in mobile and cellular radio technology—Motorola.

For more information on Motorola's **MHW926** and **MHW927A/B** power modules, request RF Products Data Sheet MHW926/D and Selector Guide SG46/D by calling 1-800-441-2447, by sending the coupon below, or by writing Motorola Semiconductor Products, P.O. Box 20912, Phoenix, AZ 85036.

MOTOROLA

To: Motorola Semiconductor Products, Inc., P.O. Box 20912, Phoenix, AZ 85036

Please send me the RF Products Data Sheet MHW926/D and Selector Guide SG46/D.

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INFO/CARD 19

tion algorithm. Each device containing the chip will have two unique "keys" that must both be used to decode messages. "Key escrow" agents will retain the keys and provide access only to government officials with legal authorization to conduct a wiretap. Companies and individuals will gain protection for proprietary and private information, while law enforcement agencies will still be able to lawfully intercept the phone conversations of criminals. NIST is working on a standard to facilitate federal government procurement and use of the "Clipper Chip."

British Association for Low Power Radio Devices — The Low Power Radio Association is a British group that was "formed to promote the responsible use of low power radio equipment and to act as a channel of communication between the Radiocommunications Agency and LPRA members." Their primary focus is on devices that operate at short range and with less than 500 mW power. Application areas include RF data communications, telemetry, video/audio, telecommand, wireless local area networks, alarms, RFID, telemonitoring, radio microphones and teleapproach. For more information about the association contact: LPRA Secretariat, The Old Vicarage, Haley Hill, Halifax, HX3 6DR, United Kingdom. Tel: (44) 0422 359161. Fax: (44) 0422 355604.

Electrical Engineering Scholarship Established — Telecommunications Techniques Corporation has established the Joseph A. Sciulli Memorial Scholarship to be awarded to a student majoring in electrical engineering entering the University of Maryland at College Park. The scholarship will be awarded to a Maryland resident entering the Engineering Department based on financial need, scholastic excellence in math and science, leadership, and work ethics. The scholarship program, including the selection process, will be administered by the University of Maryland's College of Engineering. An optional summer internship at TTC will enable the student to get hands-on experience.

Epitronics Awarded Contracts — Epitronics recently announced they were awarded an ARPA SBIR Phase I contract to investigate Indium Phosphide based device structures for microwave and millimeter wave monolithic integrated circuits and a subcontract with Ware Technical Services, Inc. to deposit strain-free epitaxial layers on indium gallium arsenide substrates.

New RF Tags Unique to Industry — Lanex Corporation recently announced the availability of an active RF tag that is unique to the industry because of its long range, low cost and ability to read multiple tags in a designated field of reception. The Locator, using state-of-the-art technology is detectable from 40 to 50 feet away from the reader and sells for $32 in small quantities. The tag system can be used for security management, inventory and process control solutions and is compatible with most host systems or can be used in a stand alone version. The Locator contains a receiver and transmitter that are powered by a miniature lithium

battery with a life expectancy of 10 years. Communications are achieved by the reader transmitting a low frequency signal and receiving a high frequency one.

Garbage Collection Goes High-Tech — The city of Edmond, Oklahoma recently began replacing garbage cans with new 105 gallon waste bins, each of which is equipped with a small radio frequency transponder from Texas Instrument TIRIS™ division. Each passive transponder contains a unique identification code, which allows electronic readers to accurately identify the bin's owner. Fully-automatic trucks lift the plastic household waste bins, identify the owner, weigh and empty the garbage, place the bin back on the sidewalk and later download the information, allowing the owner to be billed based on the garbage's weight.

Aviel Electronics Relocates — Aviel Electronics has announced its new address. They are now located at: Aviel Electronics, 5530 South Valley View, Suite 103, Las Vegas, NV 89118. Tel: (702) 739-8155, 8157. Fax: (702) 739-8161.

HK Microwave Changes Name and Relocates — HK Microwave recently announced that it has changed its name to Dynatech Spectrum, Inc. and has relocated the company from Santa Clara to Milpitas, California. The decision on the name change was made to give a closer association with the company's product lines and their markets. Their address is now 100 S. Milpitas Blvd., Milpitas, CA 95035. Tel: (408) 956-9570. Fax: (408) 956- 9595.

EEsof Forms Strategic Alliance — EEsof, Inc. has announced the formation of a long-term strategic alliance with STEP Electronics to exclusively distribute and support EEsof products throughout the Pacific Rim and other parts of the world. STEP employees are currently undergoing extensive training in the sale and support of the full range of EEsof products for workstations and PCs.

General Microwave Awarded Contract — General Microwave Corporation recently announced that it was awarded a $2.2 million contract by BMY Combat Systems for the development of a microwave Terrain Mapping Sensor. The sensor will be integrated into an Automatic Depth Control system for a tank-mounted mine clearing blade. General Microwave will use its patented short range radar technology to develop a system that provides data describing the terrain ahead of the vehicle with a high degree of accuracy.

Stanford Awarded Receiver Assembly Contract — The ASIC & Custom Products Division of Stanford Telecommunications, Inc. has announced that it has entered into an agreement with Harris Corporation's Information Systems Division to develop a specialized receiver board assembly for use in weather satellite reception. Harris plans to deploy the receiver module in a system which will receive, process, and display real-time weather satellite data.

ULTRA-BROADBAND RF POWER AMPLIFIER SYSTEMS

From KALMUS ENGINEERING, Of Course!

ALL SOLID-STATE MOS-FET RF AMPLIFIER SYSTEMS

	MODEL	RF OUTPUT	FREQUENCY RANGE	GAIN	USA PRICE
	700LC	1.5W CW	.003-1000 MHz	33dB	$ 1,695
	704FC	4W CW	.5-1000 MHz	33dB	$ 2,195
	210LC	10W CW	.008-225 MHz	40dB	$ 2,495
	710FC	10W CW	1-1000 MHz	40dB	$ 6,695
	*727LC	10W CW	.006-1000 MHz	44dB	$ 7,750
NEW	713FC	15W CW	10-1000 MHz	42dB	$ 4,250
	225LC	25W CW	.01-225 MHz	40dB	$ 3,295
	*737LC	25W CW	.01-1000 MHz	45dB	$ 9,995
	712FC	25W CW	200-1000 MHz	45dB	$ 6,950
NEW	714FC	30W CW	10-1000 MHz	45dB	$ 9,950
	250LC	50W CW	.01-225 MHz	47dB	$ 5,250
	715FC	50W CW	200-1000 MHz	47dB	$ 16,990
	707FC	50W CW	400-1000 MHz	50dB	$ 9,990
NEW	716FC	50W CW	10-1000 MHz	47dB	$ 17,950
	*747LC	50W CW	.01-1000 MHz	47dB	$ 19,500
	116FC	100W CW	.01-225 MHz	50dB	$ 8,800
	709FC	100W CW	500-1000 MHz	50dB	$ 16,990
	717FC	100W CW	200-1000 MHz	50dB	$ 19,500
NEW	718FC	100W CW	10-1000 MHz	50dB	$ 26,950
	*757LC	100W CW	.01-1000 MHz	50dB	$ 29,950
	122FC	250W CW	.01-225 MHz	55dB	$ 14,650
	723FC	300W CW	500-1000 MHz	55dB	$ 29,995
	LA500G	500W CW	500-1000 MHz	57dB	$ 53,500

RUGGED VACUUM TUBE DISTRIBUTED AMPLIFIERS

MODEL	RF OUTPUT	FREQUENCY RANGE	GAIN	USA PRICE
116C	100W CW	.01-220 MHz	50dB	$ 9,295
122C	200W CW	.01-220 MHz	53dB	$ 11,750
134C	500W CW	.01-220 MHz	57dB	$ 19,800
137C	1000W CW	.01-220 MHz	60dB	$ 27,550
140C	2000W CW	.01-220 MHz	64dB	$ 44,990

Warranty: Full 18 months all parts. Vacuum tubes 90 days.

* Indicates Dual-Band System (coaxial band switching)

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INFO/CARD 23

The World's Most Complete Line of RF Power Amplifiers

More Than 200 Standard Models to Choose From

A Diverse Market for RF Amplifiers

By Gary Breed
Editor

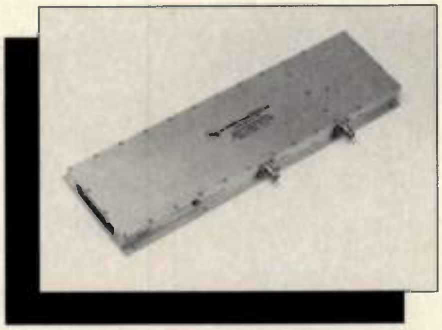
Companies making medium- and high-power RF amplifiers continue to direct their efforts toward device, design and manufacturing improvements. Current applications include new wireless communications systems in the 900 MHz-2.5 GHz range, but these applications are by no means the sole force driving amplifier development.

The observations of companies manufacturing and selling RF amplifier products suggest that the market is extremely diverse. Each company seems to have identified a specific niche that is providing a good market for its products. Rebuilding commercial communications equipment in the 400 MHz range to handle digital information was identified as a strong market by David London, Director of Marketing for Trontech. "Digital technology is coming on stream rapidly because it is the most efficient way to transfer information," he explains. On a different course is Milcom International, which recently introduced a wideband distributed amplifier providing 50 watts of linear power from 1 to 512 MHz. Among potential customers, Milcom expects this product to reach power semiconductor manufacturers who need improved test system capabilities.

The positive side of this diversity is that the overall market is not subject to the fate of a single large application. However, the negative side is that the market is hard to predict. It is difficult to decide exactly which smaller applications to target, making it hard to anticipate customers' needs. One effect is to shift amplifier sales efforts from specific products to "company capabilities." Rather than speculate on an amplifier line for a specific 1.8 GHz Personal Communications System (PCS) application, an amplifier manufacturer may choose to promote its ability to quickly design and produce amplifiers for a wide range of applications. Response to customers is more important than a catalog full of stock products.

Medium-Power Applications

Power capabilities in the 0.1 to 10 watt range are rapidly increasing in importance. Formerly reserved for drivers in higher power systems or as final amplifiers in handheld equipment, medium power amplifiers are doing a lot more, especially in receiver applications.



New applications include cellular microcells, the intended use for this Microwave Power Devices amplifier.

Dynamic range is the problem solved by using a power amplifier in a receiving system. Cellular systems, repeater sites and PCS all require receivers to perform properly in the presence of co-located transmitters. To avoid intermodulation distortion (IMD), the front-end receiver preamplifier must be capable of handling signals varying from the microwatts of received signal from a handheld cellular phone user to watts of power as the cell site transmits another conversation on a nearby channel. The transistors used must also have a low noise figure, to maintain good receiver performance. Companies like Locus, Inc., Trontech, Mini-Circuits and TIW Systems are pursuing this market area, along with several others.

Modular amplifiers for medium power are more common than they were just a few years ago. Q-Bit, Avantek, Cougar Components and Watkins-Johnson are among numerous suppliers of TO-8, connectorized, and flatpack amplifiers. The availability of these amplifiers makes it possible to speed a product's development cycle by using off-the-shelf parts to avoid the time and cost of circuits designed and built from scratch in-house.

New Devices

The search for higher efficiency, low voltage operation, lower cost and better performance (linearity, noise, etc.) has kept transistor manufacturers busy. Class A efficiency at 1 GHz, for example has improved by over 5 percent over the last few years, and class C amplifiers have seen improvements of 10 to 30 percent, depending on the frequency. Bipolar transistors, silicon FETs and GaAs FETs have all been the subject of research efforts. New buzzwords in power devices are HFET (heterostructure FET), HBT (heterojunction bipolar transistor) and PHEMT (power pseudomorphic HEMT). These technologies represent the current efforts at improving amplifier performance.

Internal matching of RF power transistors has been used extensively at microwave frequencies, but it is now seen at HF and VHF/UHF. One recent addition to this group are two product families from MicroWave Technology. Their SLAM devices (basically, these are power JFETs) include internal resistors to make 50-ohm matched, self-biased class A push-pull power amplifiers with minimal external circuitry. At higher frequencies, their new self-biased GaAs FETs provide 4 to 10 watts in the 500-2000 MHz range. They can be dropped into 50 ohm systems, making them truly amplifiers rather than transistors.

Summary

The technology and applications of RF amplifiers are, unfortunately, much too broad for a short report. This brief look has identified a few trends in amplifier markets and some of the developments in amplifier products. The key word to remember is *diversity*. The market for amplifiers is becoming one of many different specific requirements, with customers requiring a particular set of performance features rather than adapting a standard amplifier to their application. **RF**

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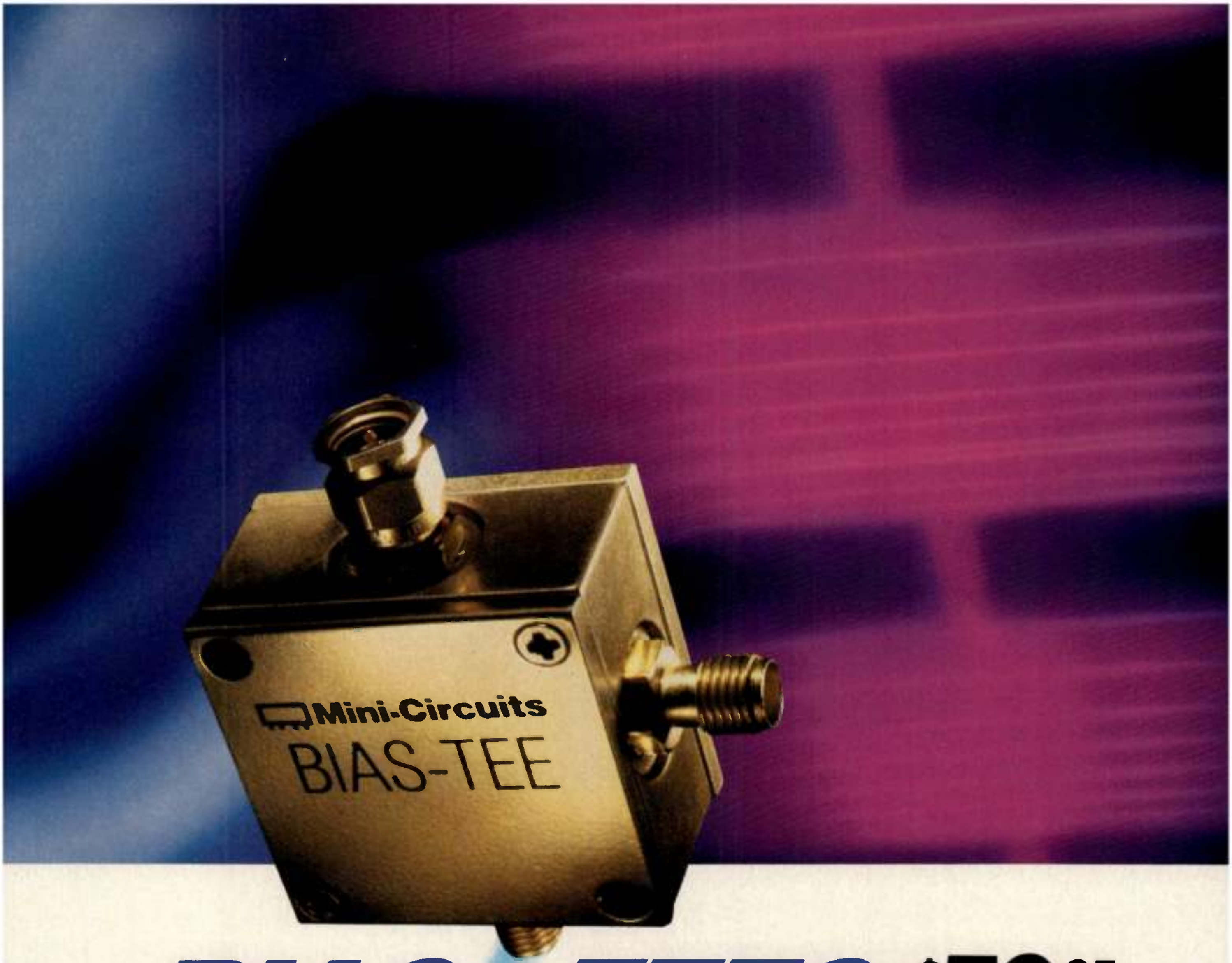

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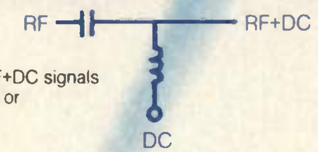

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A Low Power RF ID Transponder

By Raymond Page
Wenzel Associates

This is the Grand Prize winner in the design category of the 1993 RF Design Awards Contest. This entry exhibited both innovative use of RF technology and an elegant implementation of that technology. The author was awarded a NOISE COM model UFX-BER noise generator for bit error rate testing.

For some time railroad companies have been wrestling with the problem of tracking rail cars. This has traditionally required manual log entry of identification numbers displayed on the cars as they pass through the switching yard. Some years ago, an effort was undertaken to use an optically scanned ID system. Dirt and optical registration problems led to its demise, forcing railroad companies to revert to the manual system. RF engineers have come up with a solution, using transponders mounted on the side of the cars which are read by interrogating transceivers positioned along the track.

Design Considerations

A practical transponder design must include minimal maintenance, a rugged low profile and low cost. The most elusive of these has been low cost. Presented here is a design which meets these requirements along with a brief discussion on the current state-of-the-art in passive RF identification transponders.

An important design constraint is that the transponder require little or no maintenance. Since no power is available from the rail car, the only conventional options are batteries or solar cells that maintain rechargeable batteries. The non-rechargeable batteries require periodic replacement, and the solar cell option would be both expensive and vulnerable to the environment. A passive design eliminates the need for batteries by rectifying energy from the interrogating RF field to power the circuitry.

The harsh environment presented to an RF device mounted on the side of a rail car is a challenging problem. Minimum clearance requirements, dirt, weather, vibration and an extremely large chunk of ferro-magnetic material near the antenna have to be considered. Additionally, the unit should be encapsulated. Microstrip patch antennas have come to the rescue. They afford a low profile and can be made with an ordinary double-sided printed circuit board. The patch antenna is on the top and a ground plane is on the bottom, thereby eliminating the effects of the steel mounting surface.

A Low Cost Transponder

An unusually simple method of converting the interrogating RF field into a data-modulated signal which can be transmitted back to the reader contributes to the low manufacturing cost of this transponder design. The circuit uses only one inexpensive microwave semiconductor (a diode) and allows all parts to be mounted on an FR-4 printed circuit board with the patch antennas (Figure 1). By contrast, other approaches use expensive microwave parts, including SAW devices, oscillators, mixers, filters and amplifiers. Designs involving more RF circuitry tend to be power hungry, requiring increased RF interrogation fields.

Figure 2 shows the block diagram of the low power transponder. A 915 MHz receive antenna powers the rectifier/frequency doubler/AM modulator. It provides a rectified DC source to the MCU which returns data to be AM modulated onto the doubled frequency. An 1830 MHz antenna transmits the modulated carrier.

A reader, incorporating an unmodulated 915 MHz interrogation transmitter with low (< –60 dBc) second harmonic distortion and an 1830 MHz AM receiver, is placed a relatively short distance away from where the transponder will pass (Figure 3). The amount of transmitted RF interrogation power needed to make the system function properly at a given distance can be estimated by equation 1:

$$P_r = P_t G_t G_r \lambda^2/(4\pi R)^2 \qquad (1)$$

Where P_r is the received power, and P_t is the transmitted power radiated by an antenna of gain G_t. G_r is the gain of the receive antenna, λ is the free space wavelength and R is the distance between transmitters. G_t and G_r are the gains over an isotropic radiator. A sufficient second harmonic return path signal will occur for any combination of power gain and distance capable of energizing the MCU.

One watt of power transmitted with an antenna gain of 31.6 (16 dB) and received with an antenna gain of 2 (3 dB) allows the transponder to function from as far as 20 feet away. This suggests that just over 1 mW is adequate to energize the transponder.

The transponder's surprisingly low

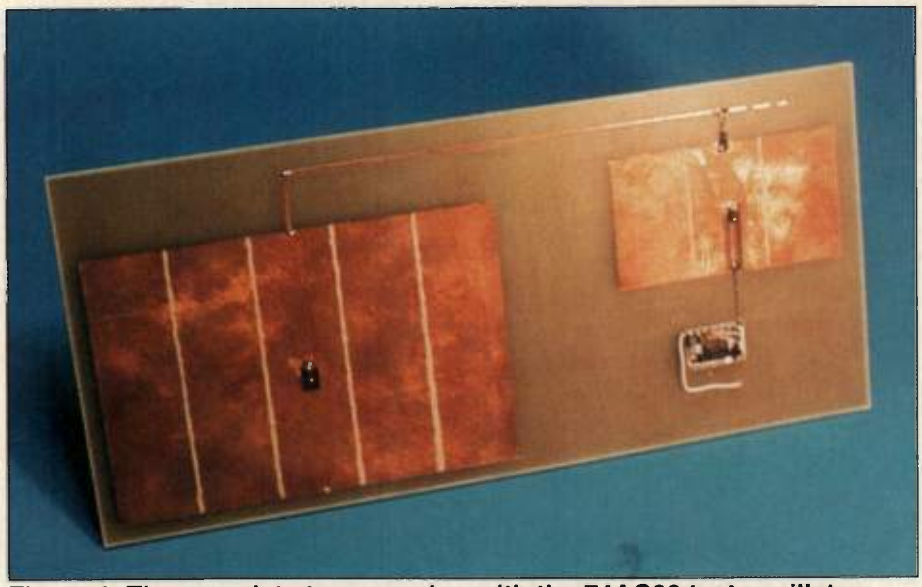

Figure 1. The complete transponder, with the 74AC00 test oscillator.

power requirement is due to its efficient means of rectification, frequency doubling and modulation. All of these functions are accomplished by a single microwave diode. A hybrid schematic in Figure 4 details the circuit. The 915 MHz patch antenna has two connections, a DC return path connected at the zero impedance point and a transmission line matched to the 120 ohm impedance at the edge of the antenna. The transmission line routes the signal to CR1 for rectification. A DC tap on the 1830 MHz antenna provides the power connection for the MCU. (See side bar on microstrip patch antennas.)

Careful placement of CR1 along the transmission line is crucial in creating the proper AC impedances for efficient frequency doubling. The 1830 MHz antenna becomes a 90 degree open stub at 915 MHz at the cathode of CR1, effectively giving the 915 MHz signal a low impedance trap to work against (Figure 5). Since the transmission line does not provide a similar low impedance on the anode side of CR1, a 90 degree open stub at 1830 MHz must be added.

Less than 100 uA are required to power the MCU (Figure 6). Consequently, little second harmonic is produced by CR1, leaving plenty of modulation headroom. Increased frequency multiplication occurs when the output port of the MCU goes low providing a path to ground for rectified current via the 1 kohm resistor, R1. Varying the value of R1 controls the modulation depth. CR2 and C2 work together to maintain sufficient voltage to the MCU while the voltage at C1 is being pulled down by the modulation action.

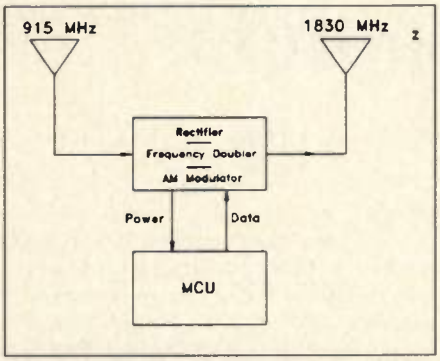

Figure 2. Block diagram of the passive transponder.

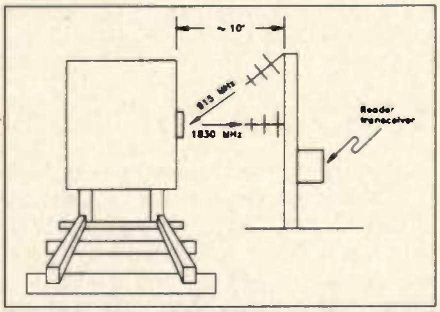

Figure 3. RF ID reader and transponder with rail car.

Performance

As previously noted, the system can operate up to 20 feet away. However, performance is measured at the 10-foot separation required during normal operation (Figure 7). For test purposes, a spectrum analyzer functions as the receiver. A 74AC00 gate oscillator in Figure 4 is substituted for the MCU to simulate load and logic level conditions. The oscillator simplifies confirmation of the concept. Three kHz modulation is used for easy detection by the analyzer.

The transponder transmits data at 94 percent AM modulation. Figure 8 is the detected 3 kHz square wave. Measurements of the rectified voltage (2.7 VDC) and current (1.45 mA DC) give 3.9 mW total power which correlates nicely with the received power (5.3 mW) predicted

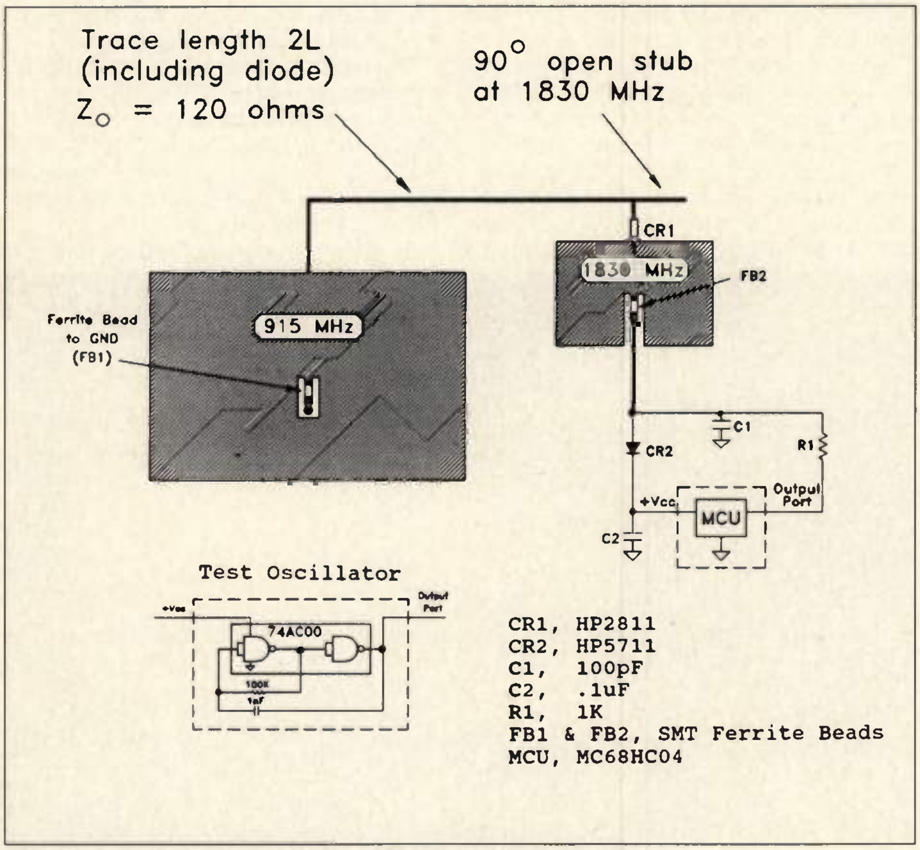

Figure 4. Hybrid schematic of transponder circuit.

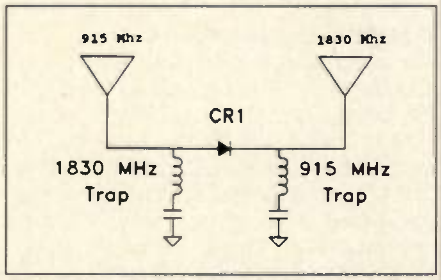

Figure 5. Equivalent AC circuit of transponder showing RF traps.

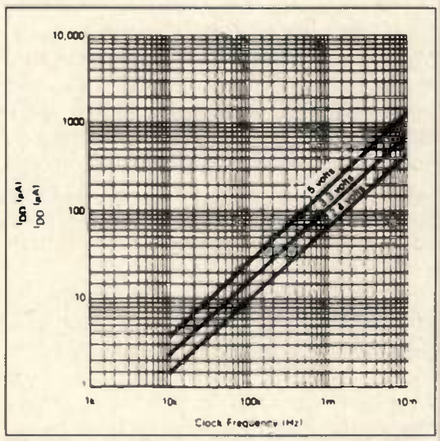

Figure 6. Current vs. clock frequency for a typical 68HC04 MCU.

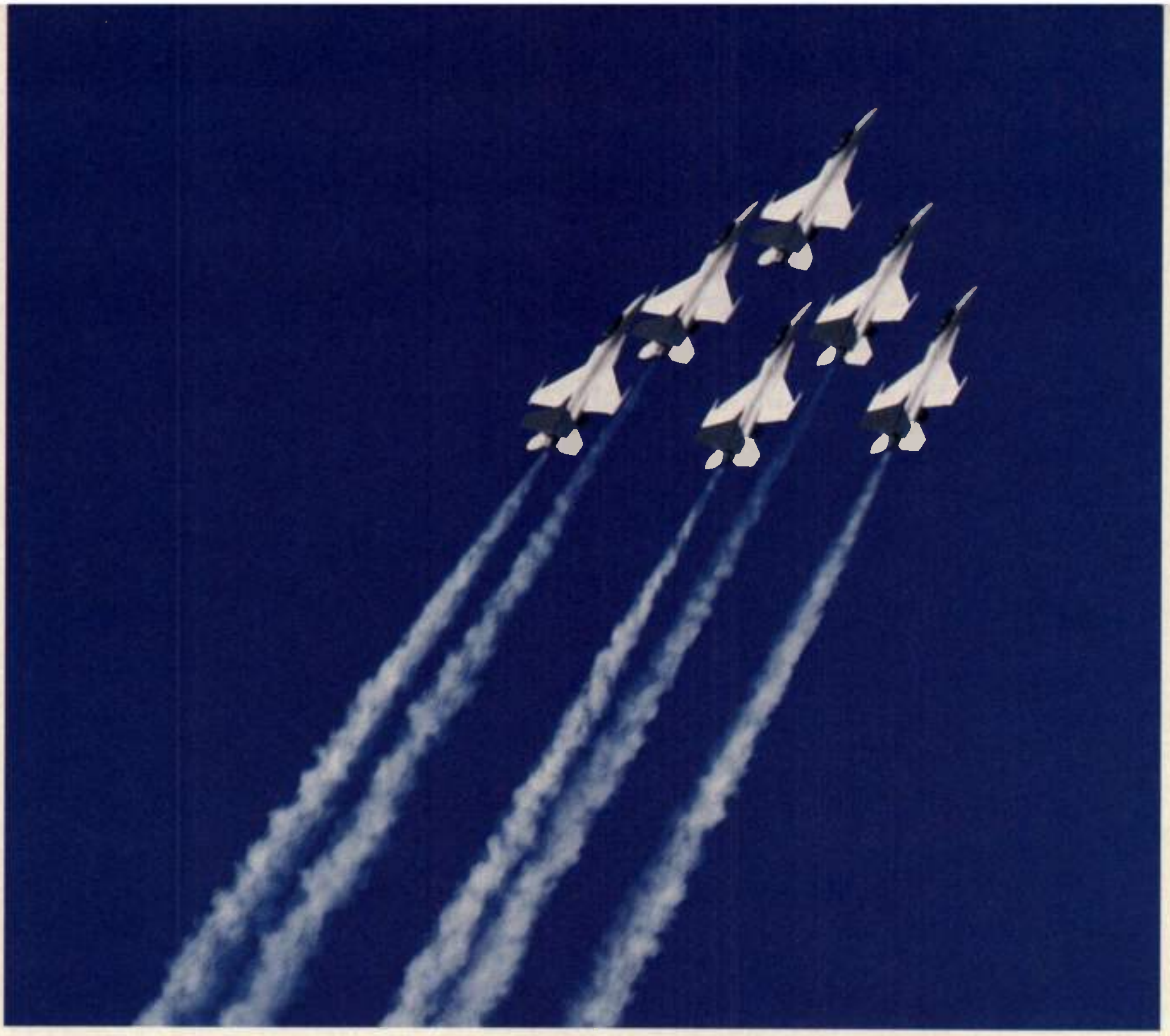

Figure 7. Test setup for transponder operating at a distance of 10 feet.

by equation 1 at a distance of 10 feet.

Improvements

Inherent compatibility with spread spectrum is provided by this design since the returned signal frequency is derived directly from the interrogation signal. Frequency spreading is limited only by the bandwidth of the patch antennas. With the simple addition of a micro-power line receiver and the associated communications software, the transponder can be upgraded for two-way information applications. Size reduction can be accomplished by increasing operating frequency at the expense of costlier substrate material. Borrowing technology from missile and aircraft radar technology the transponder could be made a part of the "skin" of its host.

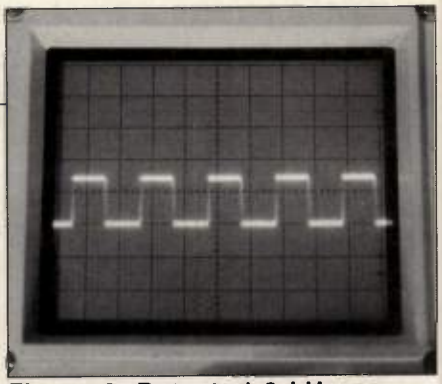

Figure 8. Detected 3 kHz square wave.

Summary

This paper has described the design, operation and application of a low-power RF identification transponder. The simple design is spectrum friendly, requiring

Rectangular Microstrip Patch Antenna

The rectangular patch antenna is essentially a resonant microstrip with an electrical length of 1/2 the wavelength of the frequency to be transmitted or received. Microstrip patch antennas work well for applications requiring a low profile, offering a height equal to the thickness of the printed circuit board from which they are made. PTFE substrates are normally used to minimize dielectric losses which affect the efficiency of patch antennas. However, FR-4 is a cost effective alternative for low power applications at frequencies below 2 GHz.

Microstrip antennas come in all sizes and shapes. A rectangular patch is chosen for its simple geometry and linear polarization when fed from the center of an edge. The input impedance varies as a function of feed location. The edge of a 1/2 wavelength antenna has an input impedance of approximately 120 ohms which drops to zero ohms as the feed point is moved inboard to the center of the antenna. This allows easy impedance matching and provides a convenient means of DC tapping the antenna as seen in the transponder design.

For simplicity, the dimensions of the microstrip patch antennas in Figure 9 are in terms of L, which is equal to 1/2 the electrical wavelength of the receive antenna (915 MHz). L can be determined by equation 2:

$$L = 0.49(\lambda/\varepsilon_R) \qquad (2)$$

where λ is the free-space wavelength and ε_R is the relative permittivity of the printed circuit board.

Bandwidth is determined by the substrate thickness and can be approximated for an SWR of less than 2 by equation 3:

$$BW = 128f^2t \qquad (3)$$

BW is in MHz, f is the operating frequency in GHz, and t is the substrate thickness in inches.

Applying equations 1 and 2 to the transponder design using 0.125 inch FR-4 substrate material with an effective permittivity of 4.7 results in a value of 2.92 inches for L and a bandwidth of 13.4 MHz at 915 MHz.

Our Design Contest Winner

Raymond Page is a design engineer for Wenzel Associates, a manufacturer of high performance crystal oscillators and frequency standards. Ray specializes in low noise designs for devices including oscillators,

phase locked frequency sources, multipliers and dividers. In addition to having fun with electronics, he enjoys outdoor sports and music. He can be reached at Wenzel Associates, 1005 La Posada Drive, Austin, Texas 78752, or by telephone at (512) 450-1400

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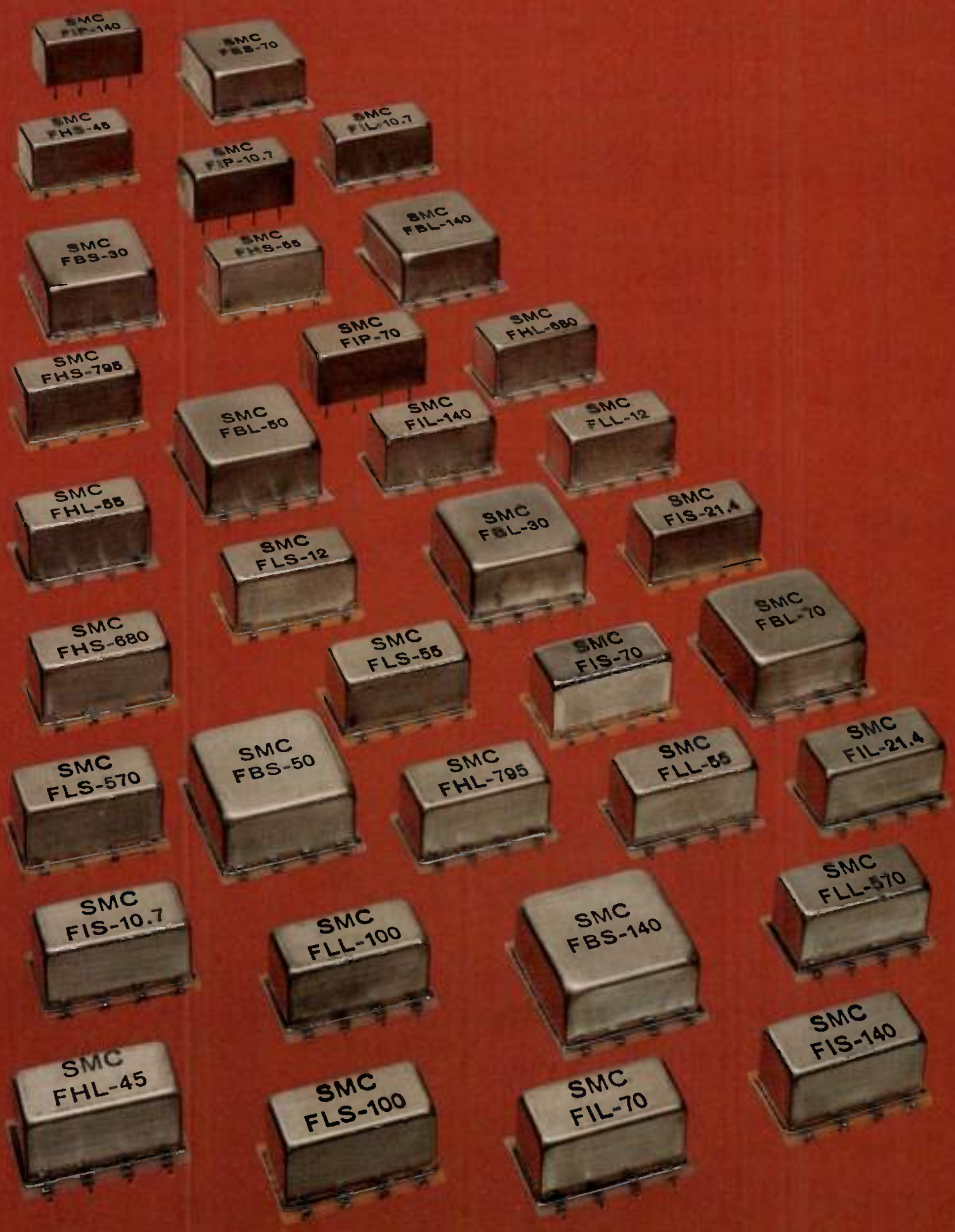

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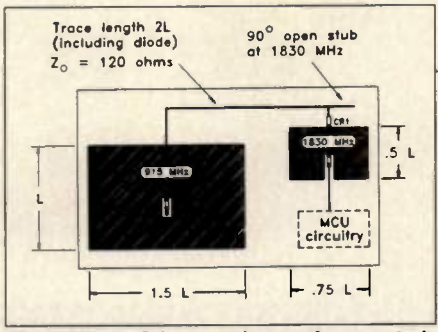

Figure 9. Dimensions for patch antennas.

minimal interrogation power and allows easy conversion to spread spectrum without modification to the transponder. Designed with one inexpensive microwave part on a single piece of FR-4 substrate, component and manufacturing costs are kept down, potentially opening up markets served exclusively by bar coding technology. Other uses include automatic tolling, inventory tracking and military vehicle security. *RF*

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1. Howard W. Sams & Co., *Reference Data for Radio Engineers*, Chapter 27, Sixth Edition, 1977.
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RF design awards

Synthesizer Design With Detailed Noise Analysis

By Terrence F. Hock
National Center for Atmospheric Research

Here is the winning software entry in the 1993 RF Design Awards Contest. The author has been awarded a software package provided by EEsof, Inc. which includes the Touchstone linear circuit analysis program for Windows and the LineCalc transmission line analysis program.

This program is designed to aid in the design of PLL synthesizers, taking into account all the noise sources present in the circuit. The program will analyze and design the loop filter and calculate the loop steady state responses, with results in both tabular and graphical form. It will also compute the reference spur levels for a digital phase/frequency detector with a differential output and can optionally add an active low pass filter to reduce the spur level. The program was written to simplify designs using commercial synthesizer integrated circuits. It operates on a DOS machine and requires a minimum of VGA graphics.

The total phase noise of the synthesizer is determined by several noise sources in the loop such as the reference oscillator, voltage controlled oscillator (VCO), phase detector and operational amplifier loop filter. Typically, the op amp noise is considered a second or third order effect when a good low noise op amp is used, but in reality the op amp may be the limiting noise factor in a synthesizer.

Program Overview

When designing a synthesizer, there are several performance criteria that must be considered: phase noise, switching speed, reference spur levels, etc. Optimizing one parameter usually gives less than adequate performance in another area. Thus, the performance of a single loop synthesizer is a compromise of all design goals.

In the past when designing synthesizers, several design tools have been used: spread sheets, small Basic programs and Spice simulators. This program was developed to combine all synthesizer design criteria into a single program which allows a design to be completed in a minimal amount of time. One parameter can be varied while the effect is observed in all other performance areas. Prior to developing this program, an adequate model for the phase noise of the synthesizer was not available.

The program calculates the noise contribution from the individual stages to obtain the total phase noise of the synthesizer. The loop bandwidth can easily be changed while observing the phase noise, which allows for easy optimization. The program displays the integrated phase jitter, providing an aid in choosing the loop bandwidth for lowest overall phase noise. The loop filter model incorporates the performance of a real op amp for the calculations of op amp filter noise, reference spur levels and steady state loop responses. The Johnson noise of the resistors is also incorporated into the op amp noise calculation.

All calculations use linear control theory in the frequency domain for both the steady state responses and phase noise calculations. A wideband synthesizer (loop bandwidth greater than 20 percent of the reference frequency) analyzed with linear analysis will introduce errors. A better approach would be to use discrete time analysis or Z-transforms for analysis. However, if wide loop bandwidths are desirable for fast switching speeds, the reference spurs may be at undesirably high levels when using the digital phase/frequency detector in a low cost synthesizer IC.

To improve the accuracy of the calculations, the user has the option of adding a delay term to the steady state loop calculations. There has been considerable discussion in past issues of *RF Design* as to what the delay value should be (1). The value of this delay term is dependent upon the type of phase detector used in the synthesizer (digital phase/frequency, sample and hold), and the location of the first pole. There is not a clear consensus on what the sampling delay should be for all cases, so its value is left for the user to choose.

Currently the program does not calculate the switching speed of the synthesizer, but uses a simple approximation for a rough estimate of the switching speed. Gavin has demonstrated how a Spice simulator can be used to estimate the switching speed incorporating all delay terms for accurate switching speed analysis (2).

Phase Noise

When designing a synthesizer for low-

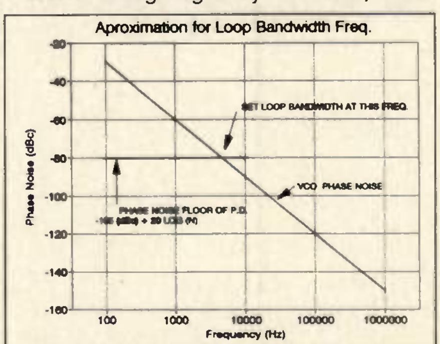


Figure 1. Method for selecting loop bandwidth.

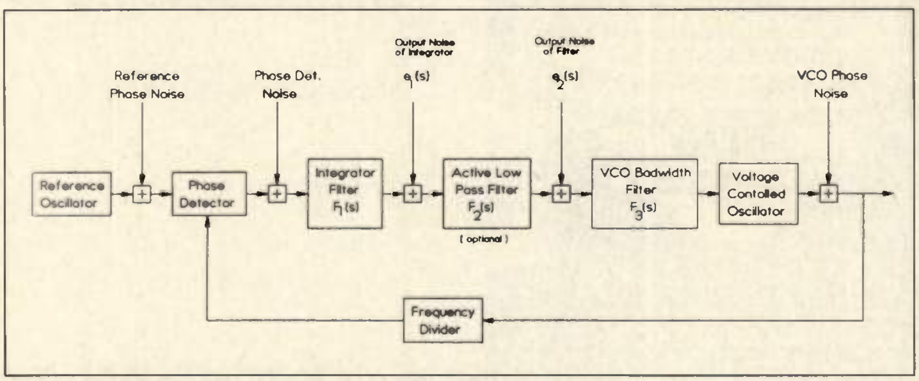
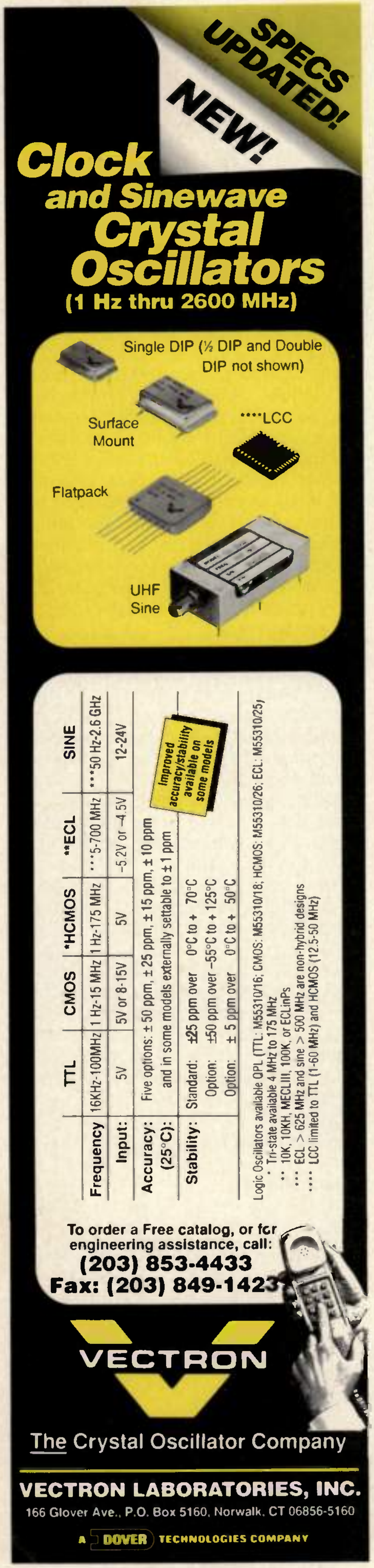


Figure 2. Noise sources in a synthesizer.



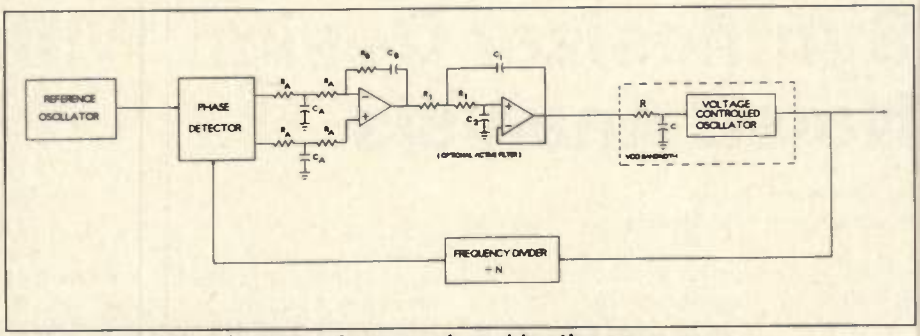


Figure 3. Synthesizer topology analyzed by the program.

est phase noise, optimum selection of the loop bandwidth is critical. Typically only the VCO noise and phase detector noise floor are considered in determining the final phase noise of the synthesizer. The phase noise inside the loop bandwidth is usually approximated by equation 1:

phase noise
= 20 log (N) + phase det. noise floor (1)

where N is the total division ratio of the VCO frequency to the reference frequency and the phase detector noise floor might be −155 dBc. The phase detector noise floor varies with the type of technology used, such as CMOS, ECL and GaAs. CMOS has the lowest noise floor of the three. For optimum phase noise performance, the loop bandwidth would be chosen where the noise floor from equation (1) intersects the phase noise of the VCO. This simplified analysis is shown in Figure 1. As a quick evaluation of where to place the loop bandwidth, this procedure is quite valid. However this method excludes the true loop behavior or closed loop response and error loop response since they modify the VCO phase noise and the phase detector noise floor.

A more detailed analysis is required to predict the actual value of the phase noise in the synthesizer. As the architecture of the synthesizer changes from

$$\text{REF OSC PHASE NOISE} = \frac{\dfrac{Kpd\ Kvco\ F_1(s)\ F_2(s)\ F_3(s)}{s}}{1 + \dfrac{Kvco\ Kpd\ F_1(s)\ F_2(s)\ F_3(s)}{sN}} \cdot \theta_{Ref}(s)$$

$$\text{PHASE DET. PHASE NOISE} = \frac{\dfrac{Kvco\ F_1(s)\ F_2(s)\ F_3(s)}{s}}{1 + \dfrac{Kvco\ Kpd\ F_1(s)\ F_2(s)\ F_3(s)}{sN}} \cdot \theta_{PD}(s)$$

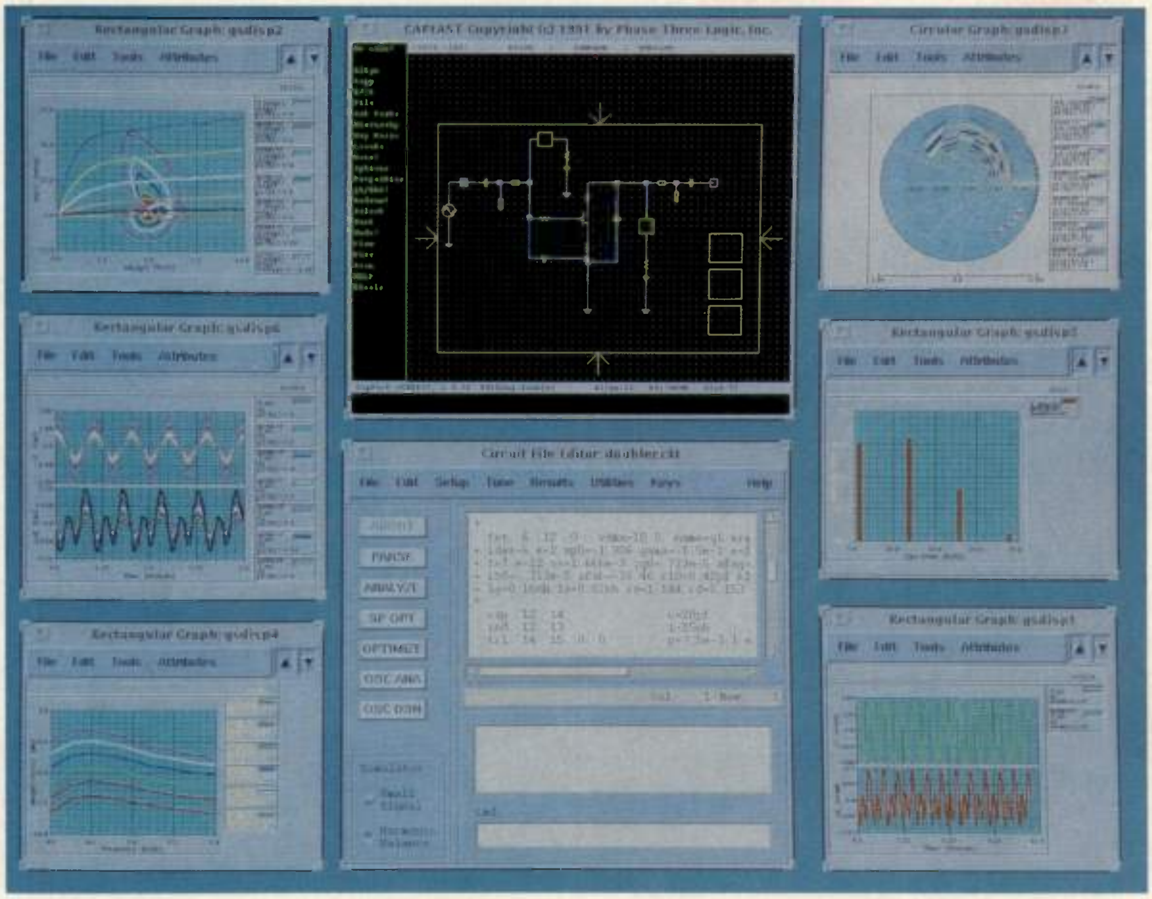
$$\text{VCO PHASE NOISE} = \frac{1}{1 + \dfrac{Kvco\ Kpd\ F_1(s)\ F_2(s)\ F_3(s)}{sN}} \cdot \theta_{VCO}(s)$$

$$\text{OP-AMP INTEGRATOR NOISE} = \frac{\dfrac{Kvco\ F_2(s)\ F_3(s)}{s}}{1 + \dfrac{Kvco\ Kpd\ F_1(s)\ F_2(s)\ F_3(s)}{sN}} \cdot E_1(s)$$

$$\text{OP-AMP LOW PASS FILTER NOISE} = \frac{\dfrac{Kvco\ F_3(s)}{s}}{1 + \dfrac{Kvco\ Kpd\ F_1(s)\ F_2(s)\ F_3(s)}{sN}} \cdot E_2(s)$$

Table 1. Phase noise system equations.

THE SMART CHOICE



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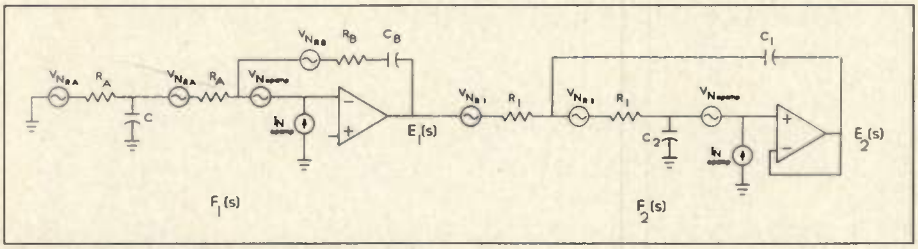

Figure 4. Loop filter noise model.

large values of N to high VCO phase noise, this simple model will have significant errors in predicting the phase noise. A block diagram of all the noise sources that contribute to the phase noise is shown in Figure 2. Equations used in the noise analysis are shown in Table 1. The op amp noise source is typically neglected when high quality low noise op amps are used in the loop filter integrator circuit. Yet, in certain cases the op amp will be the major contributor to the phase noise, even when very low noise op amps are used. This is typically the case when a VCO with a high tuning sensitivity of 25 MHz/V or greater and phase detectors with low gain (1 V/rad or less) are used.

Program Description

The program has four major sections: Design, Analysis, VCO Phase Noise and Synthesizer Phase Noise. The program assumes the user is familiar with designing synthesizers. Figure 3 is a schematic showing the circuit topology which the program evaluates. The synthesizer is a type II fifth order including the VCO bandwidth as a pole and first pole in an op amp. If an active filter is used to reduce the reference spurs, the synthesizer is a type II seventh order control system.

Design — The synthesizer design rou-

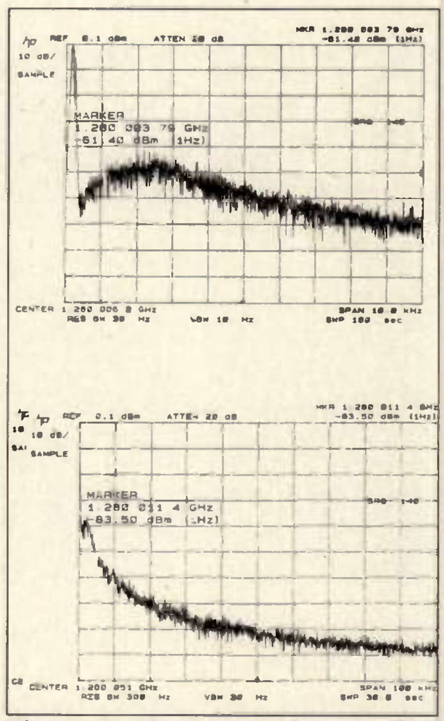

Figure 5. Measured phase noise of example synthesizer at 10 kHz and 100 kHz maximum span.

Frequency (Hz)	FILTER (dBc)	VCO (dBc)	REF (dBc)	Phase Det. (dBc)	Total (dBc)
10.0	-80.1	-84.3	-64.7	-80.8	-64.4
17.8	-80.1	-81.8	-69.6	-80.8	-68.7
31.6	-80.1	-79.3	-74.4	-80.8	-71.8
56.2	-80.1	-76.8	-79.0	-80.8	-72.9
100.0	-80.0	-74.3	-83.4	-80.8	-72.2
177.8	-79.8	-71.8	-87.3	-80.7	-70.6
316.2	-79.1	-69.5	-90.6	-80.5	-68.7
562.3	-77.7	-67.3	-92.9	-79.9	-66.7
1000.0	-75.6	-65.6	-93.9	-78.8	-65.0
1778.3	-73.4	-64.8	-94.0	-77.4	-64.1
3162.3	-73.5	-66.6	-95.6	-78.1	-65.6
5623.4	-80.6	-75.1	-104.3	-86.2	-73.8
10000.0	-89.8	-84.8	-116.1	-97.6	-83.4
17782.8	-98.8	-92.9	-128.9	-110.2	-91.8
31622.8	-108.1	-100.1	-142.8	-124.0	-99.5
56234.1	-117.8	-106.9	-157.4	-138.5	-106.5
100000.0	-127.7	-113.3	-172.3	-153.4	-113.1
177827.9	-137.7	-119.3	-187.3	-168.4	-119.2
316227.8	-147.7	-125.0	-202.6	-183.7	-125.0
562341.3	-157.7	-130.5	-218.4	-199.5	-130.5
1000000.0	-167.7	-135.8	-235.2	-216.2	-135.8

Fvco: 1280.00 MHz

Fref: 250.00 kHz

Kvco: 23.0 MHz/v

Kpd: 0.796 v/rad

Phase jitter over
100 Hz to 1 MHz
2.097 deg.

Figure 6. Program analysis of the example synthesizer (tabular form).

tine prompts the user for the common synthesizer design parameters such as VCO tuning sensitivity, phase detector gain, reference frequency, loop bandwidth, etc. The component values of the loop filter are calculated based on the equations from reference (3). The program allows the user to easily change the component values for practical sizes in the loop filter without having to reenter the initial design data. The new values are immediately updated on the screen. The program will next compute the values of the reference spurs at the VCO output. The spur calculation is only valid for a digital phase/frequency detector using differential phase detector outputs. This type of phase detector is used in the spur calculation since it has lower spur levels than the single ended output from a digital phase/frequency detector as long as the loop filter is an op amp integrator. The program then provides the option to add a 2-pole active low pass filter to reduce the spur level. The program prompts for the bandwidth and damping coefficient of the filter. The filter component values are easily changed for practical values.

The design routine will also compute and display the steady state responses of the synthesizer in table and graph form. The steady state responses computed are closed loop gain, open loop gain, error loop gain and phase margin. The open loop gain and phase margin are the most critical parameters for evaluating the loop stability. The phase margin at 0 dB open loop gain frequency should typically be 40 to 60 degrees for a well behaved loop.

PHASE NOISE COMPARISON

FREQUENCY kHz	MEASURED dBc	CALCULATED dBc
1	–66.9	–65.0
3	–61.4	–65.6
10	–83.5	–83.4
50	–106	–106.0

SWITCHING SPEED

MEASURED	1.2 msec
CALCULATED	1.54 msec

Figure 7. Comparison of measured and computed results.

Usually, a synthesizer is designed to operate at several (or many) different frequencies. In a single loop synthesizer the output frequency may vary 25 percent or more, which requires changing the divider value, which varies the loop gain. The loop should be evaluated at several operating conditions for stability and bandwidth.

Typically the tuning sensitivity of the VCO changes at different control voltages. Both the tuning sensitivity and divider values can be evaluated quickly by entering either K or F followed by new values. The output frequency is changed in lieu of changing the actual divider ratio. The loop switching speed is also calculated based on a simple approximation of t=4/loop BW.

Analysis — The synthesizer analysis routine is very similar to the design routine except actual component values of the loop integrator and active low pass filter are entered. The steady state responses of the loop are calculated along with reference spur level. The steady state responses, open, closed and error loop are calculated in table and graphic form. The switching speed is not calculated since the analysis routine does not directly compute the loop bandwidth. The analysis routine is very valuable for analyzing current designs and to evaluate component sensitivity in the loop filter versus performance variations. The output frequency and the VCO gain can both be easily changed for different operating frequencies and VCO gain varia-

Input Parameters:		Calculated Filter Parameters
Output Frequency:	400 MHz	Ra = 10.070 k ohms
Reference Frequency:	50 kHz	Rb = 57.020 k ohms
Phase Detector Gain:	.796 V/rads	Ca = 7.11 nF
VCO Gain:	5 MHz/V	Cb = 6.800 nF
Total division N:	8000	
VCO Bandwidth:	15 kHz	R1 = 5.684 k ohms
Op Amp:	LF156	C1 = 5.60 nF
Resistance Tolerance:	5 %	C2 = 1.400 nF
Loop Bandwidth:	1.3 kHz	
Phase Margin:	50 deg.	
Low Pass BW 3dB:	10 kHz	
Low Pass Damping:	.5	

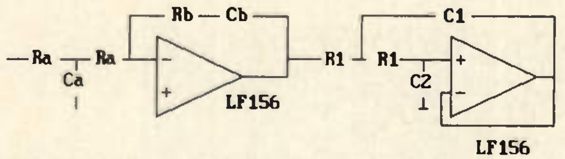

Figure 8. Input screen display.

FREQUENCY (Hz)	OPEN LOOP GAIN Gain (dB)	OPEN LOOP GAIN Angle (deg)	CLOSED LOOP Gain (dB)	ERROR LOOP Gain (dB)
10.0	74.6	1.2	70.1	-74.6
17.8	64.6	2.1	70.1	-64.6
31.6	54.6	3.7	70.1	-54.6
56.2	44.7	6.5	70.1	-44.6
100.0	34.8	11.3	70.2	-34.7
177.8	25.3	19.2	70.5	-24.9
316.2	16.6	30.2	79.2	-15.4
562.3	9.1	40.7	80.3	-6.8
1000.0	2.8	44.4	81.2	0.3
1778.3	-3.1	36.4	79.3	4.4
3162.3	-9.4	13.7	72.0	3.4
5623.4	-16.8	-28.1	62.5	1.2
10000.0	-27.7	-103.0	50.3	-0.1
17782.8	-40.4	-170.9	29.6	-0.0
31622.8	-72.4	141.0	5.6	-0.0
56234.1	-97.1	110.5	-19.0	-0.0
100000.0	-122.0	105.3	-43.9	-0.0
177827.9	-146.9	97.2	-68.9	0.0
316227.8	-171.9	91.6	-93.9	0.0
562341.3	-196.9	86.5	-118.9	0.0
1000000.0	-222.1	80.4	-144.0	0.0

50.0 kHz REFERENCE SPUR ANALYSIS

Spur level with only Op-Amp integrator [Ra,Rb,Ca,Cb].
INITIAL LEVEL: -44.2 dBc

	Attenuation @ 50.0kHz (dB)	Phase @ 1.3kHz (deg)
VCO BW:	10.0	-5.0
ACTIVE LPF:	27.0	-7.5
TOTAL	30.6	-12.5

Final Reference Spur Level:
-82.0 dBc @ 50.0 kHz

Aproximate switching speed:
3.00 msec

Figure 9. Display of steady-state responses, spur and switching speed calculation.

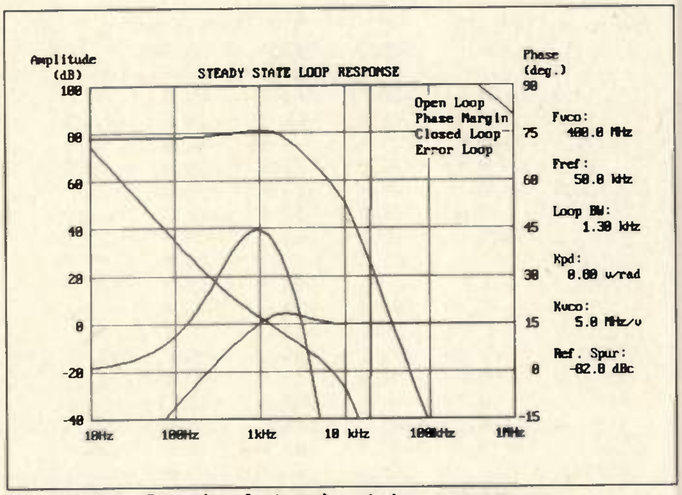

Figure 10. Graph of steady-state responses.

tions.

VCO Phase Noise — The VCO routine uses Leeson's equation for calculating the phase noise of an oscillator (4). The two dominant parameters that set the phase noise are the loaded Q of the resonator and the operating frequency or the ratio of operating frequency to Q. The output power and flicker frequency are also required in the VCO phase noise calculation. The varactor diode is also a major noise source in a voltage controlled oscillator, since it has an effective thermal noise resistance. This resistance generates a noise voltage given by Nyquist's equation which modulates the varactor diode. This noise source typically dominates the phase noise of the VCO in wide frequency tuning oscillators.

The VCO routine prompts the user for

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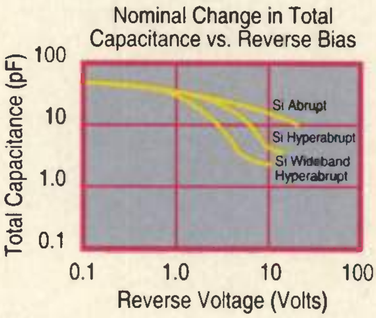

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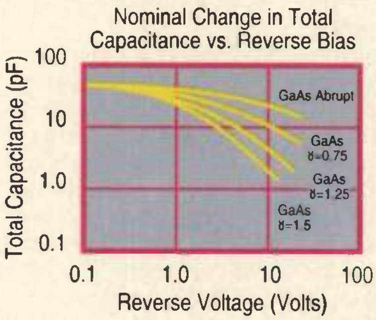
Higher Q, moderate tuning rate

Hyperabrupt (Gamma 1.5)
Wide bandwidth tuning



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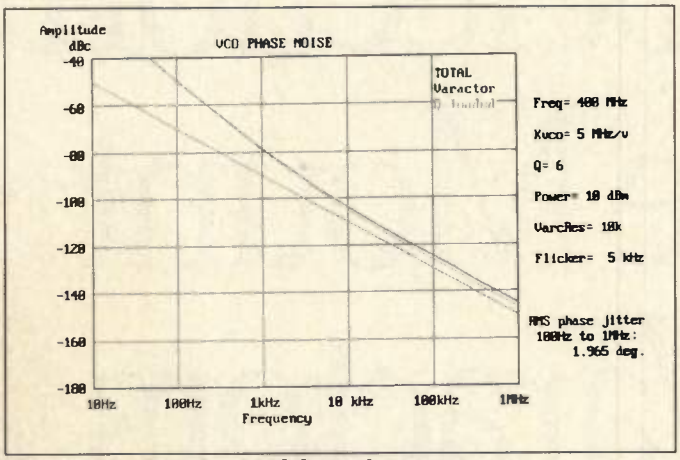

Figure 11. Graph of VCO performance.

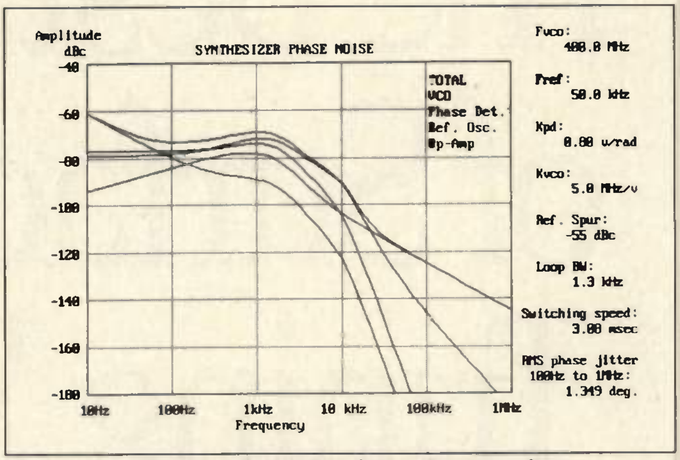

Figure 12. Graph of synthesizer phase noise.

operating frequency, loaded resonator Q, output power, flicker frequency, effective noise resistance of varactor diode and tuning sensitivity. The VCO phase noise analysis separately calculates the phase noise due to the varactor diode, Leeson's equation and the total phase noise. The performance of the VCO is easily evaluated, with the results in table and graph form.

The resonator Q and tuning sensitivity can easily be changed in 5 percent increments up or down. This feature is useful for adjusting the phase noise performance to match that of a commercial VCO. When the Q and tuning sensitivity are changed, the results are updated immediately on the screen. The phase noise from this routine will be used in the synthesizer phase noise calculations. When designing VCO's this routine is very useful for estimating the phase noise performance of an oscillator.

Synthesizer Phase Noise — The synthesizer phase noise routine is the most useful feature of the program. It computes the individual terms from each noise source in the loop and simultaneously displays the level of the reference spur, switching speed and integrated phase jitter. All design parameters are displayed on one screen while the user easily changes the loop bandwidth, observing the phase noise, spur level, switching speed, etc. The output frequency and VCO tuning sensitivity can also be changed to account for the different operating conditions as the synthesizer is set to different frequencies.

The synthesizer phase noise calculation uses data from either the Design or Analysis routine for the loop filter and data from the VCO phase noise routine. Phase noise is computed for the op amp filter, reference oscillator, VCO, phase detector and the sum of these noise sources. The op amp noise model also includes the active low pass filter if it is chosen in either the analysis or design routines. The model includes the thermal noise of all the resistors in the filters. The phase jitter is also computed by integrating the total phase noise from 100 Hz to 1 MHz. The phase jitter is useful when adjusting the loop bandwidth for lowest phase noise.

The loop bandwidth, tuning sensitivity and output frequency can all be easily changed in either the table or graphic display of the synthesizer phase noise. When a change is made, all data is updated automatically on the screen. When the loop bandwidth is changed, the feedback capacitor is held constant so all resistor values change. This is very important when the loop bandwidth is reduced as all the resistor values will increase, thus increasing the op amp noise level. The loop bandwidth cannot be changed when using data from the analysis routine. The phase noise plot has five parameters plotted simultaneously, so when a parameter like VCO gain is changed several times the graph becomes cumbersome to evaluate. A redraw feature is available by simply pressing R.

Miscellaneous Features — The program also has a Help file for operating the program, but it is assumed that the user is familiar with synthesizer design. A few common parameters such as sampling delay and resistance tolerance seldom require changing and are entered in a special utility routine. The resistance tolerance is required for the spur calculation. The spur calculation computes the common mode rejection of the op amp integrator from the op amp data and resistor values. The op amp data is in a separate ASCII file that contains the individual parameters for each op amp. This ASCII data file can easily be changed with any text editor to add, delete, or modify the op amp data file. The READ.ME file provides the format for the ASCII op amp data file. A second ASCII data file is required which stores all synthesizer data parameters as the default values when the program is terminated. These parameters will then be recalled when the program is run again. The op amp and PLL data files must reside in the same directory when the program is executed for proper operation.

General Operation

The program is designed to be very easy to use and operate with a minimal number of key strokes. At each prompt there will be a question which will ask for a value, such as phase detector gain, or a single letter designating an option. At the end of a prompt, the default value is shown in brackets. By simply pressing return, the program accepts the default value. When only one parameter has to be changed, this feature quickly allows the user to go through the input parameters without having to retype every value. An option selection will have different letters in brackets; pressing the letter will execute that command.

Printing — The results on the screen can easily be dumped to a printer using the PrintScreen key. A hard copy of the graphs can also be printed by running the DOS utility GRAPHICS.COM prior to running the program, and pressing [Shift] and [Print Screen] simultaneously. If a Laser printer is being used, be sure to add the appropriate extension to the graphics command given in a DOS

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UPC1659 600MHz to 2300MHz 23dB G_P	**UPC1675** To 2100MHz 12dB G_P	**UPC1676** To 1300MHz 20dB G_P 4.0dB NF	**UPC1677** To 1700MHz 24dB G_P P_{out} = 19.5dBm	**UPC1678** Up to 1900MHz 23dB G_P P_{out} = 18dBm	**UPC1688** Up to 1000MHz 21dB G_P 4.0dB NF

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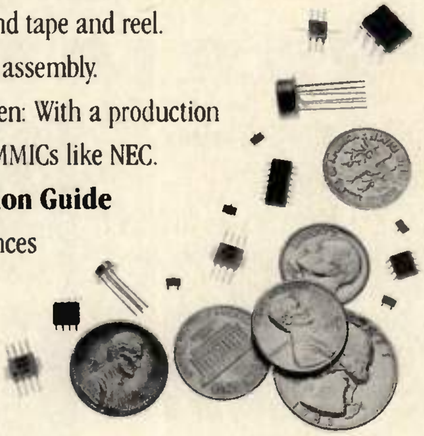

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manual.

Phase Noise Model — Several assumptions are made in the computation of the phase noise. The steady state equations used for the analysis and design routine are used for the phase noise calculations in modifying the appropriate noise source in the loop. The Z-transform would be the better choice for improved accuracy in wide loop bandwidth designed synthesizers. However for small loop bandwidths relative to the reference frequency, linear equations provide excellent results. The most unique feature of the program is that it incorporates all noise contributions from the loop's integrator and active low pass filter.

A block diagram of the noise sources considered in this program are shown in Figure 4. The noise sources $E_1(s)$ and $E_2(s)$ are lumped together into the op amp noise source when displayed on the screen. $E_1(s)$ represents the output noise of the integrator circuit including the thermal noise of all the resistors. $E_2(s)$ represents the output noise of the active low pass filter including the thermal noise of the resistors. The noise from the low pass filter is only considered if selected in the design or analysis routines.

Since most synthesizers use CMOS technology, the final frequency division of the VCO is with CMOS technology. CMOS has the lowest phase noise of all technologies available. Therefore the frequency divider phase noise was omitted from the analysis. However, the reference oscillator noise floor and divider noise are equivalent in the system equations. The phase noise of the reference oscillator is based upon Leeson's equation with a very high Q representative of a crystal oscillator. The reference oscillator usually dominates the total phase noise below 100 Hz in designs with high divider values.

An accurate model of the phase detector noise floor is not available for the various types of technologies and types of phase detectors so a constant value is entered by the user. Even though the phase noise density of the phase detector is not constant versus frequency, the noise floor will dominate the total phase noise typically at only one frequency range. The value entered should correlate to this frequency region.

The noise model of the loop filter accounts for the internal noise generated by the op amps (input noise current density and input noise voltage density) and the thermal noise of the resistors. Figure 4 shows the noise sources considered in the filter noise model used by the program. Only the Johnson or thermal noise of the resistors is considered. Contact or popcorn noise in the resistor is not considered, as this noise is very dependent on the type of resistor used and is not well defined. Metal film resistors are a good choice in the integrator filter for low noise since they have minimal contact noise. The output noise of each filter stage is computed for the calculation of the phase noise. The flicker frequency noise of the op amps, typically DC to 100 Hz is not considered in the calculation of the op amp noise. Since the phase noise of the synthesizer is typically dominated by the reference oscillator at these low frequencies, it would add little to program accuracy.

The op amp noise is minimized by keeping the input resistors of the integrator circuit small. The input noise current is effectively magnified by the size of the input resistors. Also the Johnson noise from the input resistors can become excessive if the resistors are large. If the low pass active filter is implemented, it does contribute noise to the overall loop filter and is added to the op amp noise display.

The program requests a VCO bandwidth which is used in the steady state analysis and synthesizer phase noise calculations. This is an additional pole in the loop. The VCO bandwidth in the analysis is represented as a single RC low pass filter. The VCO bandwidth can be either the actual bandwidth of the VCO or an RC low pass filter located at the input of the VCO. This additional pole filters the noise generated in the op amp circuits which can improve the phase noise performance when op amp noise is high.

Example

It is very easy to determine the dominating noise source at a given offset frequency from the carrier in a synthesizer. In loops with high division values, the phase noise inside the loop bandwidth is typically dominated by the phase detector noise floor and the loop filter (op amp noise). If a VCO with poor phase noise is used, it still will contribute significant

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noise inside the loop bandwidth as there is not enough loop gain to reduce the VCO noise. The phase noise is dominated again by the op amp noise and VCO noise outside the loop bandwidth. It becomes clear that for low phase detector gains and large VCO tuning sensitivities the loop filter is a major noise contribution to the overall performance.

The op amp noise can be minimized by using a VCO which has a lower tuning sensitivity or a phase detector with higher gain. A sample and hold phase detector with high gain of 10 V/rad or more may significantly improve the overall phase noise by minimizing the op amp noise. A second approach would be to use a phase detector with a current driver output that does not require an op amp yet is still a type II control system (5).

A synthesizer operating at 1280 MHz with a reference frequency of 250 kHz was designed and analyzed using this program. The main design goal was for low phase noise using a commercial VCO, good sideband reference suppression greater than 60 dB, and switching speeds less than 5 ms. Figure 5 is a plot of the measured phase noise of the synthesizer and Figure 6 is a plot of the calculated data from the program. Note the small noise peak at an offset of 3 kHz in the measured data, this agrees with the program's prediction. The noise peak may be mistaken for a marginally stable loop with a low damping coefficient but the frequency step response shows that the loop is well behaved. There is quite good agreement between the measured data and calculated data. A comparison of results is shown in Figure 7, along with measured switching speed for a 200 MHz change. The simple switching speed approximation in the program gives a good first order approximation.

Figures 8 through 12 show a second design example of a synthesizer, to illustrate the operation of the program and the graphical outputs.

Conclusion

Using this program, it is easy to evaluate which component in a synthesizer is the dominating noise source at a given offset frequency from the carrier. The engineer can quickly evaluate several different synthesizer architectures for best performance without prototyping. In today's competitive market, an accurate synthesizer model is extremely valuable in minimizing development time. This program has been shown to have good agreement with measured data taken from an actual synthesizer.

I would especially like to thank Mitch Randall and Tom Thomas for their valuable feedback in evaluating the program as it was developed. ***RF***

This program is available on disk from the RF Design Software Service. See page 73 for ordering information.

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Our Software Contest Winner

Terry Hock is a Senior RF Engineer in the Atmospheric Technology Division of the National Center for Atmospheric Research, where he has worked since 1982. He has BSEE and MSEE degrees from the University of Colorado-Boulder. His experience is primarily in the design of many types of communications systems. Recently, Terry was responsible for setup and support of a group of advanced weather stations, part of a multi-national research effort studying the El Niño phenomenon in the South Pacific. He can be reached at NCAR, P.O. Box 3000, Boulder, CO 80307, or at (303) 497-8767.

Contest Winners Demonstrate Innovation and Hard Work

The 1993 RF Design Awards contest is over and two RF engineers have been awarded Grand Prizes. Terry Hock is the winner of the Software Contest with a PLL analysis program, and Ray Page picks up the top prize in the Design Contest for his passive RF transponder circuit.

This year's contest, the eighth running of the RF Design Awards, was the most difficult to judge. Last year's winners, Tom Hack (design) and Don Miller (software) joined RF Design's Consulting Editor Andy Przedpelski and Editor Gary Breed on the judging team.

In the Design Contest, the judges' challenge was to compare very dissimilar entries. Different entries emphasized design methods or circuit construction, covering the RF spectrum from low frequencies to microwaves. Within this mix of applications, finding the best combination of an innovative idea, a proper engineering problem-solving approach, and good documentation was a daunting task.

Judging the Software Contest was even more difficult. The entries had to be judged on both engineering merit and implementation in a computer program. The team had to examine the philosophic issue of which aspect is most important — difficulty of the engineering problem, technical accuracy of the computations, power and easy use of the program, and overall usefulness to the RF engineering community. Finding the right balance required more than just engineering expertise.

The Software Contest

At the top of software heap is Terry Hock, an RF design engineer at the National Center for Atmospheric Research. He is awarded the Version 3.5 software package from EEsof, Inc. Terry's program covers important phase-locked loop design and analysis factors using equations from linear control theory. Noise analysis received special treatment in his program, using Leeson's equation to calculate VCO phase noise, and adding the effects of the reference, phase detector and op amp loop filter. Steady-state response, switching speed estimation, and extensive plots of PLL parameters are other major features. His complete write-up can be found on page 37 of this issue.

Terry notes that the program had an interesting period of development. Most of it was written during a research trip that included stays on various South Pacific islands, part of an international group studying the El Niño phenomenon that affects global weather patterns. Apparently, there was plenty of time available to write computer code after beachcombing and admiring the tropical flora and fauna!

Several other entries were also of extremely high quality, making judging very difficult. Among the best of the rest was a symbolic circuit analysis program by Henry Yiu of Beckman Instruments, which calculates AC or DC transfer functions in an interactive format. Henry's program is an excellent tool for obtaining a mathematical analysis of circuits.

Another software entry of note is the Tiny Electromagnetics Simulator by Jonathon Cheah of Hughes Network Systems. His program uses the Transmission Line Model to solve and plot the time domain response of a transmission line structure.

Other entries of note include a set of ECM analysis tools by Ronald Day of ITT Avionics, an S-parameter-based amplifier design aid by Dale Henkes of Philips Consumer Electronics, and a mixer analysis program by Richard Yaeger of AEL Defense Corp. These and other entries of merit will be published in RF Design over the next several months.

The Design Contest

Ray Page receives the Grand Prize of a Noise Com UFX-BER noise-based communications test instrument. Ray is an engineer at Wenzel Associates, and entered a design for a remotely-powered RFID transponder. His design recovers energy from an illuminating RF source to power a digital circuit. Meanwhile, the illuminating source is multiplied by two, and the resulting second harmonic is modulated to send data back to the interrogating unit.

Ray's circuit represents a mix of RF techniques: low-power digital circuit considerations, patch antenna design, multiplier and filtering techniques, and extensive work in the test setup to illuminate the unit and detect the re-transmitted signal. With much work in the engineering community on short-range, low-power RF systems, this design should generate plenty of interest. The design is described in Ray's article on page 31.

Design winner Ray Page admires his new UFX-BER test set from Noise Com.

Terry Hock's software victory was rewarded with an EEsof Version 3.5 package.

A few other designs deserve special mention. Pavlo Bobrek of Loral Data Systems entered a high speed, high voltage isolated line receiver that was developed to minimize the effects of lightning- induced currents. Thomas McDermott and Roy Keeney of Toshiba America MRI described the design of a programmable gain amplifier with constant phase shift, and Paul Shuch of Pennsylvania College of Technology entered a novel circuit for testing radar detector sensitivity. You will be able to examine these circuit design entries, along with several others, in the coming months.

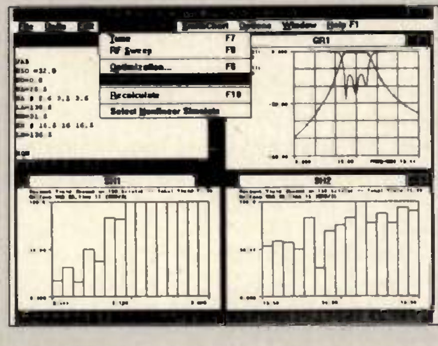
Conclusion

Once again, RF engineers have shown us that creativity and hard work go together. But, these winning entries are only a hint of the quality and depth of work that goes on in RF engineering labs around the world. Congratulations to the winners, to all those who entered, and to every other RF engineer — you all deserve recognition for your fine efforts. *RF*

The Grand Prizes!

Software Contest — Version 3.5 for Windows from EEsof, Inc.

Winner Terry Hock receives a complete design software package for the PC, operating under Windows 3.1. Version 3.5 from EEsof includes Touchstone linear analysis, Libra nonlinear simulation, and LineCalc transmission line design. This package uses the highly visual Windows environment for flexibility in graphical and physical circuit representation.



Design Contest — UFX-BER Test Instrument from Noise Com

Ray Page takes home the UFX-BER Noise Generator for bit error rate testing. Most new RF communications and control systems t·ansmit digital information, and this instrument offers advanced testing of these systems by injecting White Gaussian noise into the IF or RF of the system. The unit displays the actual Eo/No or (C+N)/N.

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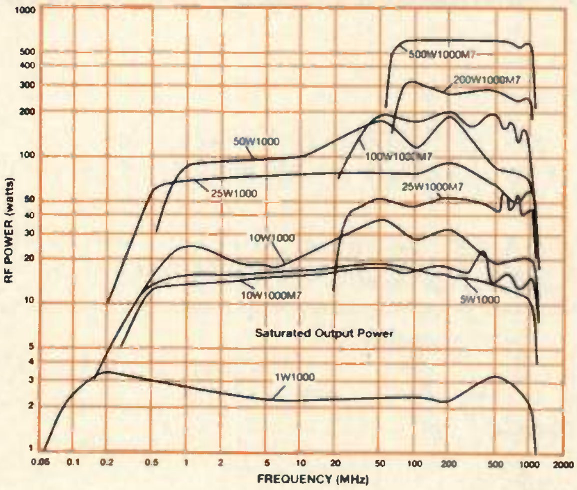

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ISBN 0-07-070964-5 | McGraw Hill | 541pg. | **$70**

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Dr. Frank Baylin, Richard Maddox, John McCormac

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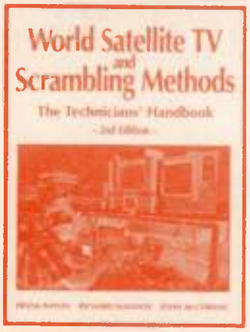
ISBN 0-917893-15-8 | Baylin Publications | 356pg. | **$40**



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Dr. Frank Baylin, Brent Gale, John McCormac

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Dr. Frank Baylin

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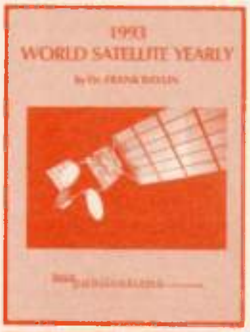

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Norm Dye, Helge Granberg

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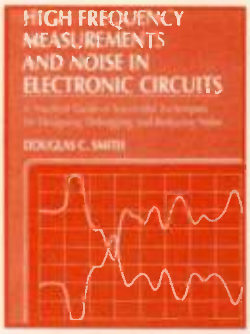

TECHNICAL BOOKS

30. The ARRL Antenna Book
Edited by Gerald L. Hall

This guide to antenna design and construction is definitely *not* just for hobbyists. It is a carefully researched collection of successful antenna designs, combined with basic antenna theory. The practical information on construction methods, materials, and various antenna configurations span the HF through microwave bands.

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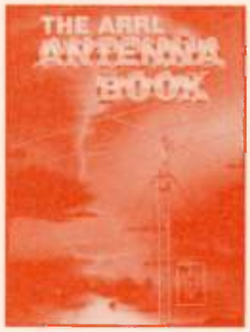

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32. Analysis and Design of Electronic Circuits Using PCs
By Greenbaum, Besser, Biehl, Pollard and Osann

Circuit simulation and synthesis using personal computers is growing rapidly in RF engineering. This book takes you step by step through the computer-aided modeling process with special attention to CAD at microwave frequencies and the use of computers in automated test and production.

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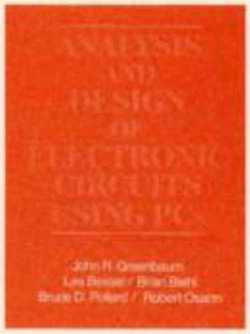

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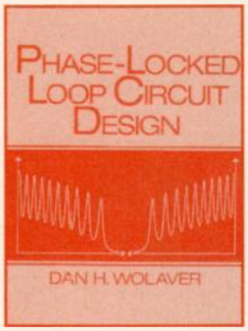

This book successfully combines theory and practice. Phase-locked loop design is presented as *circuit design,* not just as a set of theoretical equations and transfer functions. Examples given are ''real-world'' applications in signal generation, clock recovery, filtering, modulation and demodulation.

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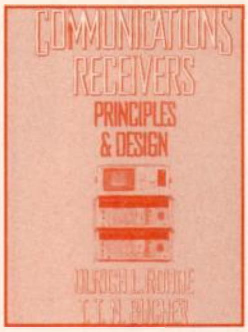
By Ulrich L. Rohde and T.T.N. Bucher



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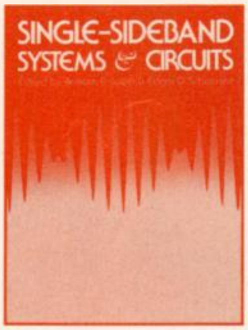

The legendary HF communications handbook from the engineers at Rockwell International/Collins Radio. Overall system design data is followed by examinations of each portion of an HF radio, including filters, mixers, oscillators, synthesizers, modulators, detectors, power amplifiers, signal processing and power supplies. This book contains information on system performance requirements that is not available anywhere else.

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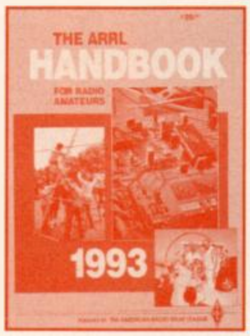

A comprehensive handbook of radio communications design and construction techniques. Analog, digital, RF and microwave basics are complemented with practical circuit examples and projects. Chapters on antennas, modulation methods and test procedures offer valuable references. Written for the advanced amateur, this book should also be on every professional's bookshelf.

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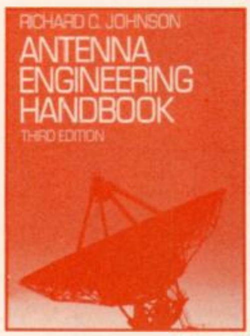

This book belongs in the reference library of every engineer involved in antenna design and development. Antenna theory and applications cover the frequency range from LF through microwaves, including communications, radar, direction-finding, satellites, aircraft, broadcasting and radio astronomy. Each chapter is authored by a leading expert in that area of antenna engineering.

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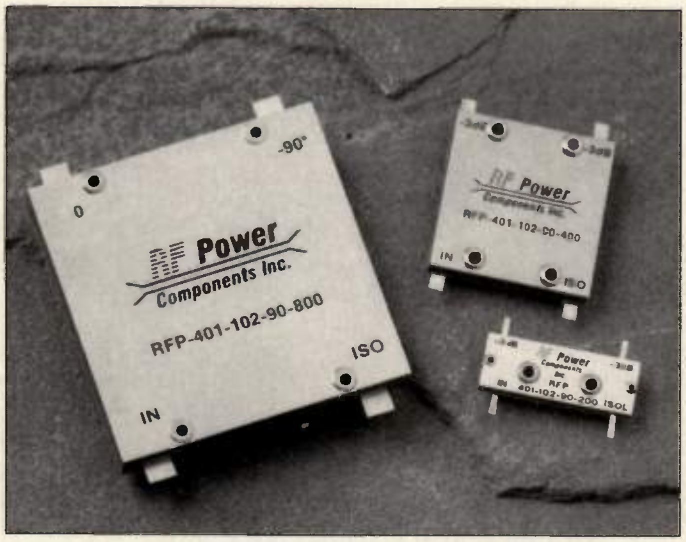
High Power Hybrid Couplers



A new family of high power ninety degree hybrid couplers which cover the 400-1000 MHz range has been introduced by RF Power Components. Designed for power combining and dividing, these models feature backward wave stripline designs with custom dielectrics. This technology allows for use in high power applications. These high power couplers are provided in three models — models RFP401-102-90-200, RFP401-102-90-400 and RFP401- 102-90-800 offer 200, 400 and 800 W CW power handling, respectively. The 400 to 1000 MHz frequency range is unique to the industry since similar competitive models typically cover the 500 to 1000 MHz range and only offer power capabilities up to 400 watts. High isolation, low insertion loss and good phase balance are additional features provided in all three devices. The couplers are available in non-magnetic drop-in versions which are ideal for medical equipment applications. Connectorized versions are also available, with SMA connectors provided in the 200 W model and type "N" connectors in the 400 and 800 W models. Sizes range from 1.35 × 0.50 × 0.12 inches for the 200 W device to 2.65 × 2.65 × 0.35 inches for the 800 W device.
RF Power Components, Inc.
INFO/CARD #250

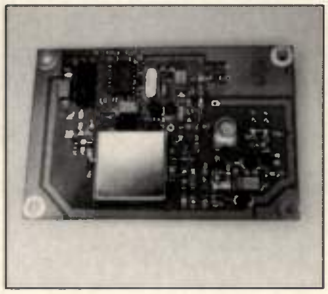
Power Grid Tubes

Svetlana Electron Devices of St. Petersburg, Russia and Huntsville, Alabama has announced a line of power grid tubes. The tubes, many of which have broadcasting applications, are exact, plug-compatible replacements for tubes currently

manufactured in the United States, Great Britain and elsewhere. The initial line of Western-style, plug-compatible tubes includes the Svetlana 4X150A, 4CX250B, 4CX250BC, 4CX350A, 4CX350AC, 4CX15000A, 4CX15000J, 5CX1500A and 5CX1500B. In addition to the plug-compatible tube types, Svetlana offers several tubes new to the Western market, including the 4CX800A and 4CX1600A, which are well suited for SSB linear amplification at 750 PEP and 1500 W CW key down.
Svetlana Electron Devices, Inc.
INFO/CARD #249

Wireless System Modules

Sciteq announces a new line of modular solutions for wireless. This new series of aggressive Sciteq products for wireless now includes an upconverter, downconverter, AGC amplifier, power module, and a wideband synthesizer. The synthesizer, the VDS-9001, is a low-cost, DRO-based synthesizer that covers all PCS bands up to 1 GHz, and supports all the common PCS step sizes of 25, 30, 50 and 100 kHz. Phase noise is excellent at –70 dBc. The interface is simple, the unit in-



cludes an onboard crystal reference, and the complete synthesizer draws only 30 mA at +5 VDC. Price of the VDS-9001 is under $100, even in modes t quantities, and availability ranges from stock to 12 weeks depending upon quantity and band selected.
Sciteq Electronics, Inc.
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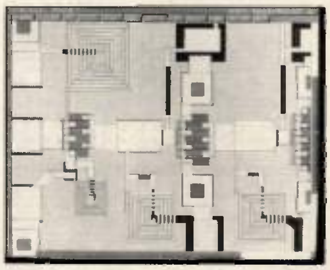
Planar Electromagnetic Simulator

Momentum from Hewlett-Packard computes the S-, Y-, or Z- parameters of arbitrary multilayer planar patterns. Patterns such as matching networks, filters, transmission-line discontinuities and other passive circuitry



found on a PC board, hybrid or MMIC design are efficiently and accurately solved with this new product. Momentum uses the method of moments for fast simulation, and its computer-generated grid composed of arbitrarily sized rectangular and triangular cells allows accurate simulation of arbitrarily shaped planar structures. Momentum allows open-boundary simulations, letting it account for radiation losses without requiring a box or enclosure. The simulator is available both as part of the HP RF or Microwave Design Systems, or as a stand-alone program with links to EEsof and Compact circuit simulators.
Hewlett-Packard Co.
INFO/CARD #247

DACs Designed for DDS

Qualcomm has introduced the Q2500 series of digital-to-analog converters (DACs). The series is well suited for use with high-speed direct digital synthesis (DDS) and other high speed applications. The Q2510 and Q2520 devices offer excellent spectral purity performance and 10-bit or 12-bit digital inputs, respectively. The 10-bit Q2510 is TTL-interface compatible and boasts update speeds to 100 Msps. The device offers fast settling time (4.5 ns to ±1/2 LSB), low power consumption (1.2 W), and low glitch impulse (1.5 pV-s). The 12-bit Q2520 updates to 80 Msps,

settles to ±1/2 LSB in 27 ns, consumes only 800 mW, and has glitch impulse of 28 pV-s. Both devices are available in 28-pin PDIP and COIC packages, operating over the –25 to +85 degree C temperature range. Prices for the Q2510 start as low as $38.10 (qty. 1000); the Q2520 starts at $34.50 (qty. 1000).
Qualcomm, Inc., VLSI Products Group
INFO/CARD #246

Product Spotlight: ICs

Receiver Front End

Philips Semiconductor announces the first integration of a low-noise amplifier, mixer and

voltage controlled oscillator in a single 3 V device. The SA620 is a complete 3 V front end solution targeted for portable cellular phones, cordless telephones, RF data links, UHF frequency conversion and spread spectrum receivers. The LNA exhibits a 1.6 dB noise figure and 12 dB power gain at 900 MHz. The active mixer supplies 6 dB power gain with an 8 dB noise figure. The SA620 consumes 8.8 mA at 3 V and is enclosed in the SSOP 20 package. The device is available now and is priced at $5.22 in quantities of 1000.
Philips Semiconductor
INFO/CARD #245

Viterbi Decoder ASIC

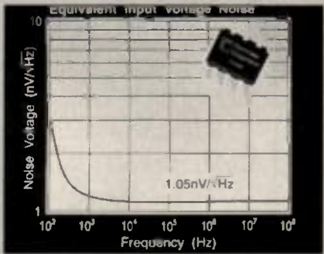
Stanford Telecom's ASIC and Custom Products Division announces the STEL-2060, a Viterbi decoder operating at data rates up to 45 Mbps. The STEL-2060 operates at constraint length K=7, has multiple coding rates, internal depuncturing capability, differential decoding, an invert G2 descrambler, and an internal BER monitor and counter. The STEL-2060 is currently available and is priced less than $20 in commercial volume quantities.
Stanford Telecom
INFO/CARD #244

Front-End Circuit

GEC Plessey has released its latest 1 GHz radio receiver front-end circuit, the SL6444. The device integrates a low-noise amplifier and double balanced mixer in a single 14-pin small outline package. The function blocks may be used independently. The SL6444 operates from 2.7 V supplies, consumes only 2 mA, has 19 dB gain, 3 dB noise figure and third order intercept point of –12 dBm. It is available in a 14-pin small outline package and is priced at $3.18 in 1000s.
GEC Plessey Semiconductors
INFO/CARD #243

Low Noise Op Amp

Comlinear's CLC425 combines a wide-gain bandwidth (1.7 GHz) with an ultra-low input noise (1.05 $nV/\sqrt{Hz}$, 1.6 $pA/\sqrt{Hz}$)



and excellent DC characteristics (100 uV vos, 2 uV/degree C drift). The CLC425 employs a traditional voltage-feedback topology and operates from a ±5 V supply.
Comlinear Corp.
INFO/CARD #242

Prescalers

Motorola has introduced seven new prescaler products offering toggle frequency ranges of 1.1, 2.0 and 2.8 GHz, a variety of divide by ratios, low power consumption and supply voltages of 2.7 to 5.0 VDC or 4.5 to 5.5 VDC. Many devices are available in both 8-pin DIP and SOIC packages. Suggested resale pricing ranges from $4.03 to $6.52 in 1000-piece quantities.
Motorola Semiconductor
INFO/CARD #241

DC-Restored Video Amp

The EL4089C is an 8-pin DC-restored monolithic amp subsystem that includes a high quality video amplifier and a DC nulling sample-and-hold amplifier that will stabilize video performance. The device has a 70 MHz bandwidth, 0.02 percent differential gain and 0.1 degree differential phase. The EL4089C is available in an 8-pin P-DIP and 8-lead small outline packages at $5.75 and $5.85 respectively in 100-unit quantities.
Elantec, Inc.
INFO/CARD #240

GSM Amplifier

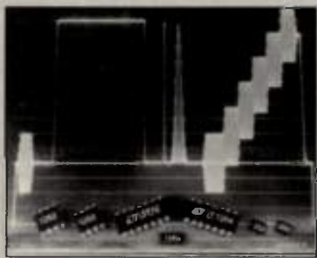
The GSM Low Noise Amplifier from AT&T Microelectronics is a thin film, balanced amplifier design. It operates in the 890 to 915 MHz frequency range and exhibits exceptionally low noise (1.3 dB max.) and high third order intercept (38 dBm). While tailored for the GSM band, it provides similar performance in the 824 to 849 MHz AMPS band.
AT&T Microelectronics
INFO/CARD #239

Transfer Switch

Mini-Circuits has introduced the model MSWT-4-20 GaAs transfer switch. The surface mount device offers repeatability (typically 4.5 sigma from mean), wide bandwidth (DC-2000 MHz), low video leakage to 50 ohm ports (typ. 15 mVpp at 500 MHz video BW), and very fast switching (4 ns typ.). Pricing for the MSWT-4-20 is just $5.75 each for quantities of 1 to 9 units.
Mini-Circuits
INFO/CARD #238

Video Op Amps

Linear Technology offers the LT1252/53/54 family of low cost, 100 MHz video amplifiers. The LT1252 is a single amplifier for $1.75; the LT1253 is a dual for $2.49, and the LT1254 is a quad for $4.49. These current feedback amplifiers operate over supply voltages from single 5 V to ±12 V.
Linear Technology Corp.
INFO/CARD #237



SEMICONDUCTORS

Power Modules

Microwave Technology is expanding its "SLAM" line of ultra-linear, class A power modules operating in the HF band. The SLAM-0122 is a self-biased, push-pull device that operates at 50 W with 14 dB typical gain. The SLAM-0133 is a single-ended device with 10 W power output and 13.5 gain.
Microwave Technology, Inc.
INFO/CARD #236

Face Bondable Diode

Alpha Industries introduces its new Metal Electrode Leadless Face Bondable (MELF) diode. The rectangular surface mount package lends itself to high volume manufacture and has no wire bonding internally. The device is capable of three times greater power handling than round MELFS with tungsten plug vs. copper dumet.
Alpha Industries, Inc.
INFO/CARD #235

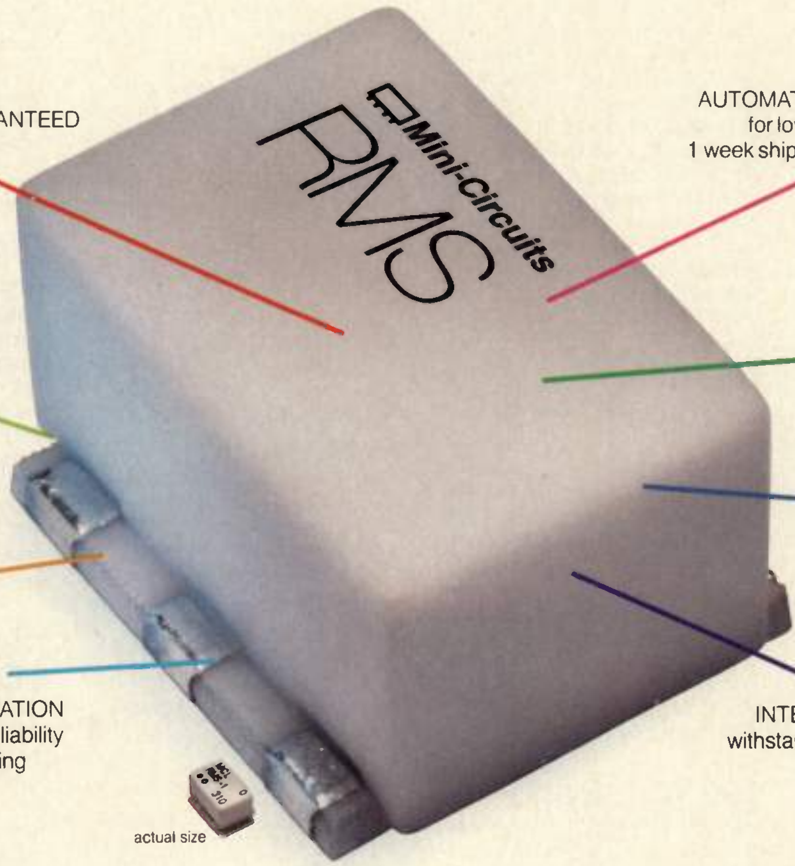
Broadband Detectors

Metelics introduces a line of detectors that offer high sensitivity and low VSWR. The detectors are available in a number of packages. Depending on the model, the frequency range is 2 to 8 GHz or 8 to 18 GHz.
Metelics Corp.
INFO/CARD #234

Power FETs

Available from California Eastern Labs are high efficiency, L-band, S-band and UHF power FETs. The 20 W NES1417-20B

MIXERS UNPRECEDENTED IN VALUE.



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Model	LO (dBm)	Freq. (MHz) LO,RF	IF	Midband, dB Conv. Loss	Isol L-R	L-I	$ ea. (10-49)
RMS-11X	+7	5-1900	5-1000	7.1	29	31	3.95
RMS-11F	+7	350-2000	DC-400	5.5	31	30	4.95
RMS-30	+7	200-3000	DC-1000	6.5	26	22	6.95
RMS-25MH	+13	5-2500	5-1500	7.5	32	32	7.95

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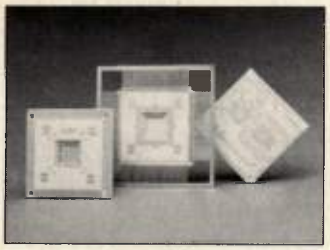
covers the 1.4 to 1.7 GHz range; the 20 W NES1818-20B covers 1.7 to 1.95 GHz applications; the 10 W NE345L-10B covers L- and S-band and UHF frequencies.
California Eastern Laboratories, Inc.
INFO/CARD #233

TOOLS, MATERIALS AND MANUFACTURING

Ceramic Packages

Diacon's line of multilayer ceramic packages features 50 ohm controlled impedance signal lines and power supply de-coupling. The MMIC packages are designed for surface mount applications and feature direct contact heat sinking and useful frequency response to 12 GHz.



Diacon
INFO/CARD #231

Manufacturing Services

American Sub-Assembly Producers (ASAP) has broadened its multi-faceted facility by offering complete kitting to shipping of high quality components and subassemblies. Among the services offered are: coil winding, hybrids, mechanical/electro-mechanical assembly and RF testing. ASAP is completely compliant with MIL Spec. 45208 and thoroughly ESD protected. All manufacturing personnel are certified to MIL Class C assembly requirements.
American Sub-Assembly Producers
INFO/CARD #230

Tuning Wrench

Coilcraft's Trituner™ tuning wrench is a unique 3-in-1 tool for adjusting variable components. The non-magnetic tool has one end stepped with a hex tip for 1/428 cores and a smaller hex tip for 10-32 cores. The other end has a molded-in ceramic straight tip for 6-32 slotted cores. Trituners cost $1.39 each in a pack of 12 dozen.
Coilcraft
INFO/CARD #229

AMPLIFIERS

Cellular Microcell Amplifiers

Microwave Power Devices has developed a multi-channel, cellular microcell amplifier. The class A amplifier, model LWA 880-30/14288, operates over 869-894 MHz with linear output power of 44 dBm (at 1 dB gain compression). Third order intercept is 54 dBm. Nominal gain is 30 dB with ±3 dB manual control. Attenuation in 1 dB steps over a 30 dB range is TTL controlled.
Microwave Power Devices, Inc.
INFO/CARD #228

Blanking Amplifier

A high-speed blanking RF power amplifier from LCF Enterprises has an output power level of 100 W CW and covers the 225 to 400 MHz range with 40 dB gain. The amplifier is available in a 19 inch rack-mount air-cooled system for $3950 each, with blanking and built-in protection against thermal overload and overdrive.
LCF Enterprises
INFO/CARD #227

100 W, L-Band

Power Systems Technology's model CHC178198-250/2974 is designed and packaged for mounting on the external structure of a satellite earth station antenna. Covering 1750 to 1850 MHz, the amplifier offers 250 W min. class C output power. Input VSWR is 1.3:1 and the output is circulator protected to infinity for any phase.
Power Systems Technology, Inc.
INFO/CARD #226

Compact Amplifiers

Microwave Solutions introduces broadband amplifiers with a 120 VAC power supply. Model MSH-3494201-PS operates from 0.1 to 3.0 GHz. Typical output power is +10 dBm and gain is 26 dB, with other gain and power options available. With power supply, this amplifier measures 5.4 × 3.0 × 1.9 inches.
Microwave Solutions, Inc.
INFO/CARD #225

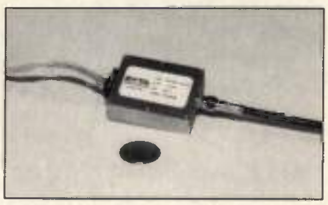
Low Noise

A low noise amplifier, model ANR 17835, operates over the cellular frequency range of 825 to 845 MHz and gain of 17 dB ±1.0 dB with flatness across the band of ±0.25 dB. Noise figure is specified at 2.2 dB max. Output power at 1 dB gain compression is 0 dB min. Input power requirements are +15 to +30 VDC at 100 mA max. Size is 0.5 × 0.9 × 0.38 inches.
TRM, Inc.
INFO/CARD #224

SIGNAL SOURCES

Miniature Synthesizer

Nova Engineering introduces a line of miniature, low power frequency synthesizers. The NS-series synthesizers are fully integrated modules, including internal TCXO, prescaler, VCO and loop filter — no external components are required. The series covers 100 to 1100 MHz, with any model tuning in a 1.5:1 range in 20 or 25 kHz steps. All models in the series share a common 2.00 × 1.50 × 0.625 inch enclosure.



Nova Engineering, Inc.
INFO/CARD #223

Low Noise Crystal Source

Techtrol Cyclonetics' 2000A-SC series ovenized crystal source features low phase noise and small size. The unit is available at frequencies between 40 and 125 MHz, meeting frequency stabilities up to $\pm 1 \times 10^{-7}$. The oscillator meets an improved phase noise level of –152 dBc at 1 kHz and a floor of –169 dBc at 100 kHz offset. Harmonics are down at –26 dBc. The unit is housed in a circuit board mountable 2.0 × 2.0 × 1.0 inch package.
Techtrol Cyclonetics, Inc.
INFO/CARD #221

SMT VCXO

The K1526 series of surface mount VCXOs is available within the frequency range of 2.0 to 33.0 MHz. Deviation sensitivity is ±50 ppm/V, with other deviations soon available. Control range is centered at 2.5 V and goes from 0.5 to 4.5 V. Temperature stability is ±50 ppm from –40 to +85 degrees C. The package measures 0.560 × 0.360 × 0.160 inches.
Champion Technologies, Inc.
INFO/CARD #220

Digitally Tuned Oscillators

NCI Systems has introduced a series of multi-octave DTOs covering the frequency range from 2 GHz to 18 GHz. The units feature fast tuning speed, low post-tuning drift, low phase noise, low power consumption and temperature stabilization. Spurious levels are typically –60 dBc.
NCI Systems, a Division of Communications Techniques
INFO/CARD #219

High Frequency ECL

SaRonix has introduced a line of ECL high frequency crystal controlled oscillators, available from 70 to 600 MHz. Both positive and negative supply voltage parts are available. The devices are available in 4- or 8-pin hermetically sealed 14-pin DIP compatible metal cases.
SaRonix
INFO/CARD #218

DISCRETE COMPONENTS

Chip Inductors

Toko America has expanded its LL2012F series of ceramic multi-layer chip inductors, adding four new values. Inductors with values of 3.3, 2.7, 2.2 and 1.8 nH were added. Members of the se-

ries have a footprint of 2.0 × 1.2 mm and a height of less than 1 mm. Typical Qs are over 40 at 600 MHz, with self resonant frequencies of up to 6 GHz.
Toko America, Inc.
INFO/CARD #217

Non-Magnetic Trim Cap

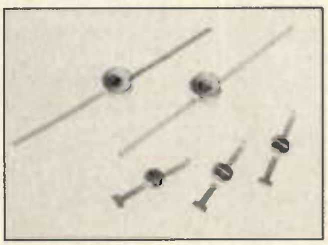
A line of non-magnetic, high-Q variable capacitors have a diameter of 0.275 inches while maintaining a high voltage rating. The capacitance spans 1 to 35 pF with a 2 kV peak test voltage.
Polyflon Co.
INFO/CARD #216

Quartz Crystals

Mercury United Electronics is introducing the UM-1, UM-5 and UM-4 series of quartz crystals. The UM-1 is 8 mm high, the UM-5 6 mm high, and the UM-4 is 4.7 mm high. The UM-1 SLIM is 2.6 mm thick versus the UM-1's 3.2 mm thickness. All crystals cover frequency from 10 MHz fundamental mode to 160 MHz 5th overtone. Frequency stability is ±10 ppm over −10 to +60 degrees C.
Mercury United Electronics, Inc.
INFO/CARD #214

Miniature Trimmers

Voltronics is producing the RP 0.094 inch and RD 0.14 inch diameter one-half turn ceramic trimmers. These designs are stable because of a compact inter-



nally supported structure. The RP tunes from 5.5 to 20 pF and the RD tunes from 5 to 25 pF. Costs are $0.58 and $0.50 in quantities of 1000 and $0.44 and $0.34 in quantities of 10,000.
Voltronics Corp.
INFO/CARD #213

Small Trim Caps

AVX/Kyocera's CTZ2 and CTZ3 trimmer capacitors have maximum dimensions of 3.0 mm and 4.5 mm, respectively. Both trimmer capacitors feature high capacitance in a small size and low profile. The trimmers are usable with reflow soldering and are washable.
AVX Corp
INFO/CARD #212

TEST EQUIPMENT

Electroptic Spectrum Analyzer

Photonic Systems now offers two Bragg-cell-based spectrum analyzers. The AOS-1000 has 1 GHz instantaneous bandwidth and 1000 frequency channels. The AOS-1000X4 has 4 GHz instantaneous bandwidth and 4000 frequency channels. Both instruments have 1 MHz channel bandwidth.
Photonic Systems, Inc.
INFO/CARD #211

RF Exposure Probes

Narda has introduced two probes shaped in accordance with the new IEEE C95.1-1991 standard for human exposure to RF/microwave energy. Models 8722B and 8732 monitor electric and magnetic fields, respectively, and have a detection sensitivity that mirrors the new standard.
Loral Microwave - Narda
INFO/CARD #210

Antenna/ Transmission Line Testing

Using one of Direct Conversion Technique's scalar network analyzers and their newly introduced DI-8 data interface, frequency swept measurements such as return loss can be quickly made on site and stored to disk. In addition, FFT capability allows sweep information to be shown in the time domain, yielding distance-to-fault information.
Direct Conversion Technique, Inc.
INFO/CARD #209

RF Component Modeling for CAD/CAE

By Les Besser
Besser Associates

This article offers a tutorial review of lumped passive RLC component models, including parasitic effects and losses. For simplicity, only first order resonances are considered, although second-order effects are also discussed. Most of the material presented here is taken from a one-week long continuing education course entitled "RF/High Speed Circuit Components: Measurements, Models and Data Extraction" (1).

During the 1970s and 80s defense-supported R&D created significant progress in the modeling of active microwave devices and various forms of physical transmission lines. At the same time, RF passive component models attracted little attention. Researchers felt that passive lumped components were very "simple" and their modeling would pose little challenge. As a result, RF components were represented by their ideal forms in early CAE programs. Recent cutbacks in military microwave spending prompted the CAE industry to pay attention to lumped-element components for commercial RF products. While we are still far from having libraries with dependable lumped models, RF component modeling is becoming an increasingly important effort and should play a more significant part in modern CAE.

Q of Physical Inductors and Capacitors

Before looking at equivalent circuits, let's define an important circuit parameter. The quality (Q) factor of a reactive component is defined by the ratio of stored energy over dissipated energy that can be expressed in series or parallel form:

$$Q = Q_S = \frac{X_S}{R_{SE}} = Q_P = \frac{R_{PE}}{X_P} \qquad (1)$$

where:

X_S is the equivalent series reactance of the series equivalent circuit.

X_P is the equivalent parallel reactance of the parallel equivalent circuit.

R_{SE} is the total equivalent series resistance of the series equivalent circuit.

R_{PE} is the total equivalent parallel resistance of the parallel equivalent circuit.

The parallel and series circuits shown in Figure 1 represent a series/parallel conversion at a specified frequency, maintaining the input Q.

RLC Component Models

Up to 1 GHz, relatively simple lumped equivalent circuits may be used to accurately describe the behavior of RLC components in chip and discrete forms. Above that frequency, depending on the size of the components, modeling may require more detailed circuits and distributed (transmission line) elements. However, most surface mount components can be adequately described by the lumped models, even up to 2 GHz.

Resistors — Figure 2 illustrates the equivalent circuit of a typical resistor, using lumped circuit elements, assuming that parasitic self-inductance and capacitance may be combined into single components L_S and C_P. Depending on the physical form of the resistor, the nominal resistance, R_N, may or may not be frequency dependent. For example, if the thickness of the resist layer does not exceed skin depth, the resistance is virtually independent of frequency. The second resistor of the model R_S represents the lead and contact resistances; they may need to be separated from the nominal resistance due to different temperature and frequency dependencies.

For low value resistors the series inductance is the prime parasitic, meaning that the total impedance of the part increases with frequency. At some point the parallel capacitance creates a low-Q parallel resonance with the inductance of the series branch, from that frequency the terminal impedance of the part declines. Below the resonant frequency the physical resistor behaves inductively while above resonance it looks like a lossy capacitor. If the nominal resistance is relatively high (several hundred ohms), the series inductance may be neglected and the parallel capacitor represents the prime parasitic. In this case the impedance declines as the frequency increases; approximating the behavior of a parallel RC (see Figure 3). While the parallel capacitance may be relatively small, i.e. a fraction of a picofarad, its effect may need to be considered. For example, if the resistor is used as a collector-to-base DC bias resistor, even a small parallel capacitance forms a significant feedback path at RF.

Inductors — Figure 4 shows the first order approximation of an RF inductor equivalent circuit. In this model the series resistance R_S represents frequency dependent ohmic self-resistance of the inductor which forms a relatively high Q series equivalent circuit with the nominal inductance. The parallel capacitor represents the inter-winding and terminal capacitances lumped together into a single

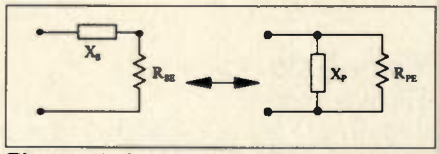


Figure 1. Lossy reactive elements may be represented by these equivalent circuits. Q may be computed either way by taking the appropriate ratios shown in equation 1.

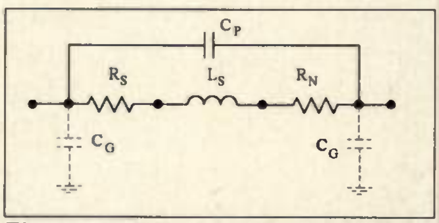


Figure 2. Lumped R_F equivalent circuit of a resistor. R_S and R_N may both be frequency dependent. L_S represents the total parasitic self-inductance and C_P the self-capacitance. If the resistor is realized above a ground-plane, the additional stray capacitances C_G also exist.

DC-2000 MHz

AMPLIFIERS

In plastic and ceramic packages, for low-cost solutions to dozens of application requirements, select Mini-Circuits' flatpack or surface-mount wideband monolithic amplifiers. For example, cascade three MAR-2 monolithic amplifiers and end up with a 25dB gain, 0.3 to 2000MHz amplifier for less than $4.50. Design values and circuit board layout available on request.

It's just as easy to create an amplifier that meets other specific needs, whether it be low noise, high gain, or medium power. Select from Mini-Circuits' wide assortment of models (see Chart), sketch a simple interconnect layout, and the design is done. Each model is characterized with S parameter data included in our 740-page RF/IF Designers' Handbook.

All Mini-Circuits' amplifiers feature tight unit-to-unit repeatability, high reliability, a one-year guarantee, tape and reel packaging, off-the-shelf availability, with prices starting at 99 cents.

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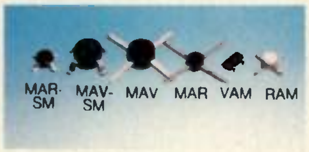


Models above shown actual size

from **99¢** Unit price $ (25 qty)

PLASTIC SURFACE-MOUNT			++VAM-3 1.45		+VAM-6 1.29	++VAM-7 1.75		
add suffix SM to model no. (ex. MAR-ISM)	MAR-1 1.04	MAR-2 1.40	MAR-3 1.50	MAR-4 1.60	MAR-6 1.34	MAR-7 1.80	MAR-8 1.75	
	MAV-1 1.15	+MAV-2 1.45	+MAV-3 1.55	MAV-4 1.65				MAV-11 2.15
CERAMIC SURFACE-MOUNT	RAM-1 4.95	RAM-2 4.95	RAM-3 4.95	RAM-4 4.95	RAM-6 4.95	RAM-7 4.95	RAM-8 4.95	
PLASTIC FLAT-PACK	MAV-1 1.10	+MAV-2 1.40	+MAV-3 1.50	+MAV-4 1.60				MAV-11 2.10
	MAR-1 0.99	MAR-2 1.35	MAR-3 1.45	MAR-4 1.55	MAR-6 1.29	MAR-7 1.75	MAR-8 1.70	
Freq.MHz,DC to	1000	2000	2000	1000	2000	2000	1000	1000
Gain, dB at 100MHz	18.5	12.5	12.5	8.3	20	13.5	32.5	12.7
Output Pwr. +dBm	1.5	4.5	10.0	12.5	2.0	5.5	12.5	17.5
NF, dB	5.5	6.5	6.0	6.5	3.0	5.0	3.3	3.6

Notes: + Frequency range DC-1500MHz ++ Gain 1/2 dB less than shown

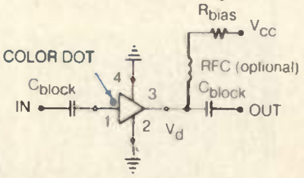
designer's amplifier kits
DAK-2: 5 of each MAR-model (35 pcs) only $59.95
DAK-2SM: 5 of each MAR-SM model (35 pcs) only $61.95
DAK-3: 3 of each MAR, MAR-SM, MAV-11, MAV-11SM (48 pcs) $74.95

designer's chip capacitor kit
KCAP-1: 50 of 17 values, 10pf to 0.1μf (850 pc), $99.95

chip coupling capacitors at .12¢ each (50 min.)

Size (mils)	Value
80 x 50	10, 22, 47, 68, 100, 220, 470, 680 pf
80 x 50	1000, 2200, 4700, 6800, 10,000 pf
120 x 60	.022, .047, .068, .1μf

Typical Circuit Arrangement



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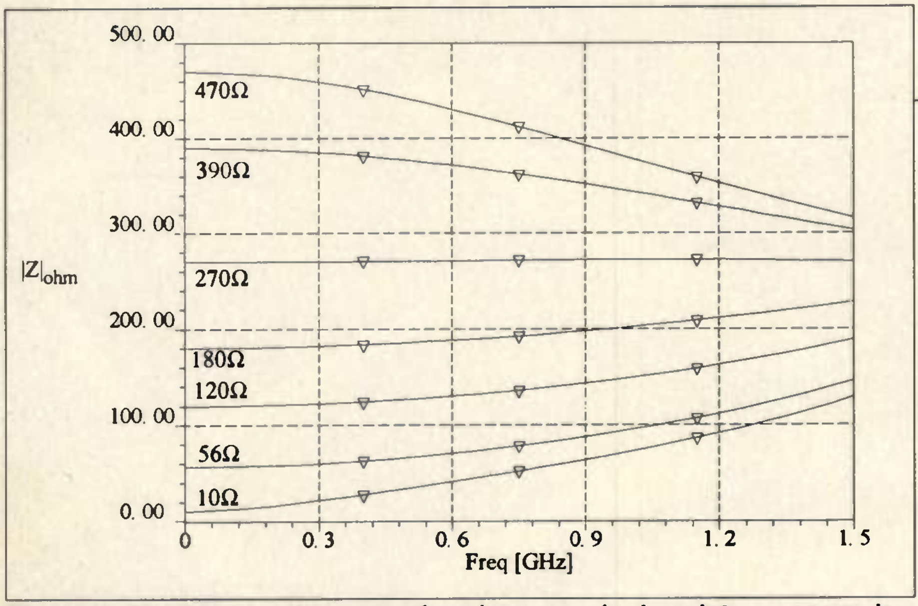


Figure 3. Impedance changes of various nominal resistors, assuming constant parasitics for all resistor values. Although the 270 ohm resistor seems to be the least frequency dependent when the magnitude of Z is plotted, the picture may be deceiving. A better way is to look at the RE(Z) vs. frequency.

part — something that is acceptable as long as the physical length of the inductor is much less than the wave length of the applicable frequency. As in the resistor model, the shunt capacitor causes parallel resonance, except the resonance is now a relatively high-Q kind (see Figure 5 for various inductor values).

The parallel resistor, R_P, of the inductor model represents the frequency dependent core losses for magnetic core type of inductors. Since permeability is a complex quantity, the frequency behavior of this resistor needs to be described very carefully, often requiring the use of piece-wise approximation or curve-fitting techniques.

The physical inductor behaves inductively until resonance; above that it is dominated by the parallel capacitor. However, in multi-layer inductors the situation is more complicated since each layer has individual equivalent circuits. In such a case, after a particular layer's parallel resonance, that layer behaves as a capacitor and eventually series resonates with another layer that still acts as an inductance, forming series resonance. Then the next layer goes through parallel resonance and so on. These parallel and series resonances are called secondary resonances and vary the terminal impedance of the inductor between high and low values, instead of just a parallel resonance and capacitive behavior of the single layer coil. If such inductor is used as a choke, assuming constant high impedance behavior, the overall circuit response might be quite different than expected.

At resonance, the inductor behaves like a parallel RLC network. The two reactances cancel each other, leaving only the resistor. To find the value of the resistor R_{PE}, the series RL branch must first be transformed to parallel form changing R_S to a high value parallel resistance R_{SP} that is equal to

$$R_{SP} = \left(Q_{RL}^2 + 1\right)R_S \qquad (2)$$

where Q_{RL} is the Quality factor of the series RL branch,

$$Q_{RL} = \frac{2\pi f L_N}{R_S} \qquad (3)$$

Combining R_P and R_{SP} provides the total equivalent parallel resistance R_{PE}. After the transformation the new inductor value

$$L_{SP} = \left(\frac{1}{Q_{RL}^2} + 1\right)L_N \qquad (4)$$

Note that for high-Q inductors $R_{SP} \approx Q_{RL} R_S$ and $L_{SP} \approx L_N$.

For example, an inductor without a magnetic core ($R_P = \infty$), having Q of 100 and series RF resistance of 1 ohm, the transformed parallel resistance at resonance is about 10 kohm. (When a magnetic core is not used, the parallel loss term R_P equals to infinity, so $R_{SP} = R_{PE}$ is the total parallel resistance.)

Capacitors — Just like the magnetic-core inductor, the capacitor also has two different types of losses, but here one is dielectric loss and the other one is ohmic type. The former is generally represented by a frequency-dependent par-

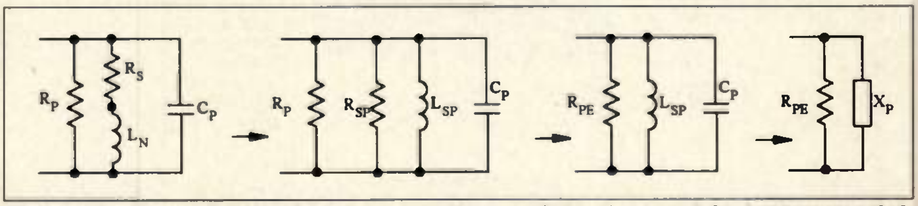


Figure 4. Four different representations of the lumped inductor model. L_N is the nominal low-frequency inductance, R_P and R_S are the frequency dependent losses (combined into R_{PE}), C_P is the self-capacitance. For Q calculation at any single frequency the basic model may be transformed to a simpler equivalent circuit shown at right.

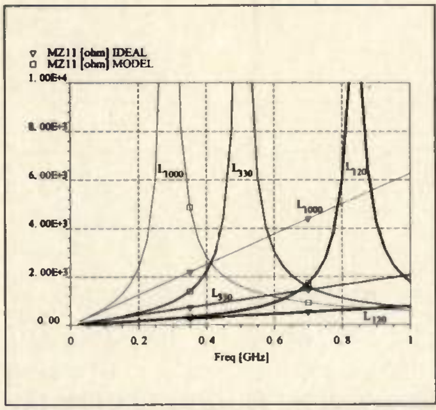


Figure 5. Impedance changes of three different physical inductors ranging from 120 nH to 1000 nH, assuming constant parasitics for all values. Straight lines correspond to ideal and curves to physical inductors.

allel resistor across the capacitance, while the ohmic losses are shown as frequency dependent series type. For maximum accuracy the ohmic losses of different origins (contact, fingers, leads, etc.) should also be separated since their frequency and temperature dependencies may differ.

The equivalent circuit shown in Figure 6 represents the first order approximation of the capacitor model. The parallel RC (representing the dielectric portion) may be converted to a frequency-dependent series form to show the series (first order) resonance of the capacitor. In this case, the converted capacitor C_{PS} resonates with the self-inductance to form a low-impedance series resonance, at which point only the combined resistance R_S plus R_{PS} exists in the circuit (see Figure 7). The sum of these two resistors is often referred to as ESR (effective series resistance), a frequency dependent term. Although ESR represents quite a low value, typically in the order of the few hundred to few tenths of an ohm, it must not be ignored in many

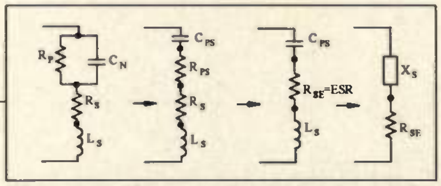


Figure 6. Lumped equivalent circuits of a capacitor. The dielectric and ohmic losses (R_P and R_S) may be combined at any frequency to a single resistor R_{SE}=ESR.

cases. In tuned circuits and high-power applications even such low resistance have significant effects. The series-to-parallel transformation changes the R_P and C_N to new values

$$R_{PS} = \frac{R_P}{1+Q_{RC}^2} \quad (5)$$

$$\text{and } C_{PS} = \left(\frac{Q_{RC}^2}{1+Q_{RC}^2}\right)C_N \quad (6)$$

$$\text{where } Q_{RC} = 2\pi f C_N R_P \quad (7)$$

The self-inductance of the contacts between the parallel fingers cause additional resonances, but these are parallel kinds, leading to high impedances. Depending on the way the capacitor is attached to the PC board these secondary resonances can be pushed to higher frequencies by the proper orientation of the chip (see Figure 8). Unfortunately, the form factors of the most commonly available parts are not appropriate to take advantage of the optimum mounting orientation. Some of the capacitor makers recognize this problem and produce "cube-like" forms to minimize the secondary resonance effects.

Parasitic Effects on Apparent Component Values

The effects of parasitic self-inductance and self-capacitance are frequently misjudged, even by those experienced in RF circuitry. For example, most designers feel that series self-inductance of a capacitor decreases the apparent capacitance, and the parallel self-capacitance of an inductor decreases inductance, while the opposite is true for both cases. The true capacitance and inductance is *always* greater than the nominal value, all the way up to the first primary resonance. For example, a 22 pF chip capacitor with 1 nH total self-inductance behaves as a 28 pF at 500 MHz, which is a 28 percent increase from the nominal value. The same is true for inductors; a 0.33 μH inductor with 0.3 pF stray parallel capacitance acts like a 1.58 μH inductor at 450 MHz, or about 380 percent increase above the nominal value.

The effects of self parasitics may be compensated by choosing *smaller* nominal values than specified. However, compensation is only possible for narrow (±5 percent) bandwidths, or even less as the frequency approaches self-resonance. In broadband applications several smaller nominal values are needed instead of a single component. A careful simulation is always recommended to prevent painful surprises. As a rule-of-thumb, the ideal and physical components behave alike until 10-15 percent of the resonant frequency. Significant changes may take place at frequencies higher than 25-30 percent of resonance.

Parasitic Effect on Q

The presence of inductive or capacitive parasitics *always* decrease the component Q, even if parasitic elements were completely lossless. To explain the reason, let's look at two cases; one for inductors and one for capacitors.

Self-inductance of a capacitor — Once we accept that self-inductance always increases the apparent capacitor value (by reducing capacitive reactance), it is easy to see that in a series RC circuit the reactance is decreased, the Q also decreases ($Q=X_S/R_{SE}$).

To illustrate this point, let's look at the 22 pF capacitor mentioned above with a 1 nH self inductance (for simplicity assume that this self-inductance comes in an ideal lossless form). We showed that the series inductance increases the true capacitance increases to 28 pF at 1 GHz, which leads to a 28 percent reduction of series reactance and Q. This situation gets much worse as we approach resonance where the effective component Q, as we define it, goes to zero.

Self-capacitance of an inductor — Since the parasitic capacitor acts as a parallel element, its effect is viewed in

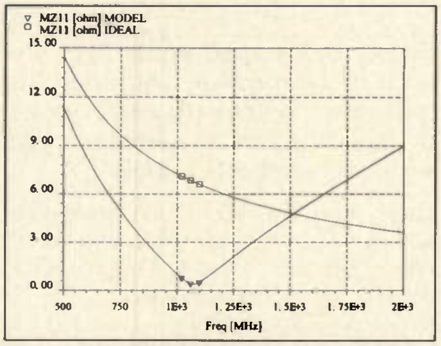


Figure 7. Impedance plot of ideal and physical 22 pF capacitors. The physical component has 1 nH self-inductance that causes a primary series resonance at 1074 MHz.

the parallel form. When the inductor is reduced to a parallel RLC equivalent circuit it was said that the presence of the parasitic capacitance *increases* inductive reactance. Parallel Q is expressed as parallel resistance over parallel reactance, therefore when reactance increases, the Q decreases.

Q Behavior of Physical Capacitors and Inductors

Even though the capacitor and inductor models contain both series and parallel circuit combinations, at a given frequency these sections may always be reduced to a single series or parallel form of one resistor and one reactive element as shown in Figure 1. Q is then computed by taking the appropriate ratio of these elements.

The frequency dependent Q behavior of inductors and capacitors are quite different. Physical capacitors have extremely high Q at low frequencies since the equivalent series reactance is very large and the series resistance (loss) is very low. As the frequency increases, the effective series resistance (ESR) increases while the series capacitive reactance decreases, resulting in a rapidly decreasing Q versus frequency.

Inductor Qs, on the other hand, behave quite differently. At low frequency the series reactance is very low, therefore Q is low. Series reactance increases with frequency, also increasing Q linearly. As the skin-effect becomes noticeable, the equivalent series ohmic resistance increases with the square-root of frequency, thereby the Q-slope changes from linear to a square-root increase. Next, the self capacitance comes into the picture, further decreasing the Q. At this point the Q levels off

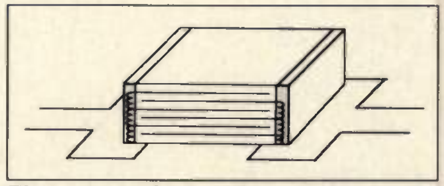


Figure 8. PC board mounting of a typical chip capacitor, showing that current flow to the top fingers must pass through additional inductance resulting in secondary resonances. These resonances may be "pushed" to higher frequencies by mounting the chip rotated 90 degrees so that the shaded surfaces contact the circuit board.

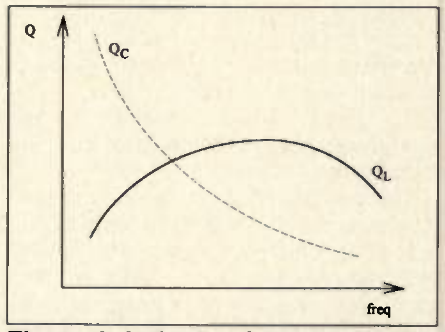


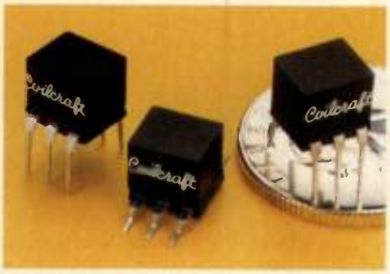
Figure 9. Inductor Q increases linearly at low frequencies, but adversely effected by skin-effect, self-capacitance, core and dielectric losses. Capacitor Q always decreases with frequency since losses increase and series reactance decreases.

with frequency, and when wire insulation and/or core losses make the picture worse, the Q begins to decline. Figure 9 compares the Q changes of inductors and capacitors.

Component Measurements

RLC components may be modeled based on one- or two-port scattering parameter measurements. For inductors and capacitors, measuring the relatively small equivalent series loss resistances may be quite tricky in one-port form. Two-port measurements on the other hand provide the necessary loss information with much higher resolution, although system impedance sometimes needs to be transformed by addition of small series capacitors.

In most cases, the components to be measured cannot be attached directly to the calibrated network analyzer; they must be placed into test fixtures. There-



Coilcraft offers low-cost wideband RF transformers

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The fixtures accommodate body sizes 0805 to 1812 and have exceptionally low stray capacitance and residual inductance. Several versions are offered for use with network analyzers, spectrum analyzers, and RF impedance analyzers such as the Hewlett Packard 4191A. Prices range from $425 to $498 depending on configuration.

For more information, contact **Coilcraft**, 1102 Silver Lake Road, Cary IL 60013. 708/639-6400.

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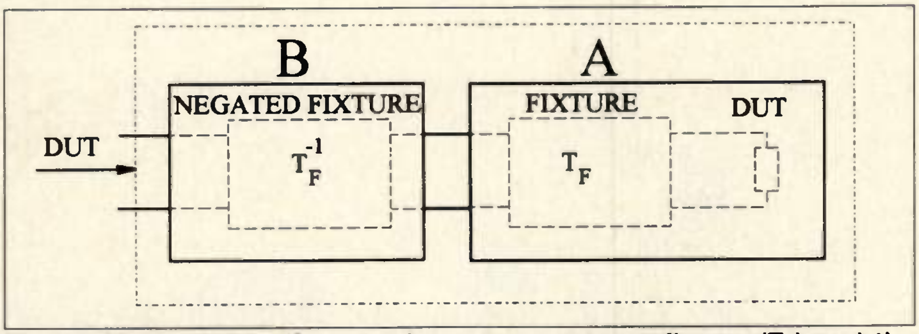

Figure 10. One-port "A" contains a two-port test fixture (T_F) and the one-port device under test (DUT). If we cascade two-port "B" that holds the "negated" equivalent of the fixture ($1/T_F$), at the input port of B we can observe DUT. Most RF CAE programs handle this procedure automatically through special negating codes.

fore, the measured data includes the parameters of the fixture and the device under test. By careful characterization, the test fixture parameters (or equivalent circuit) may be established and later de-embedded from the measured data, leaving the net parameters of the device under test. The final data then can be used for model optimization to obtain values of the component equivalent circuit. This process is illustrated in Figure 10.

Summary

With greater availability of computer simulators, it is possible to predict the behavior of RLC circuits, minimizing costly and time consuming prototype cycles. Unfortunately, circuit simulators can only predict the outcome of the circuit based on the component models provided to them. The old saying, "junk in, junk out" strongly applies in this case. Proper use of carefully selected test fixtures, calibration, measurements and data extraction can lead to appropriate models to behavior of circuits throughout the RF range *RF*

References

1. L. Besser, "RF/High Speed Circuit Components: Measurements, Models and Data Extraction," Besser Associates, 1993.

About the Author

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Data Converters Meet RF Needs Bit by Bit

By Andy Kellett
Technical Editor

The advantages of moving RF data between the analog and digital domains have probably been obvious since the first Analog to Digital Converters (ADCs) and Digital to Analog Converters (DACs) were made. However, the implementation of such systems has been anything but obvious. It has been a slow evolution from expensive converters operating at audio frequencies to relatively inexpensive devices operating at tens of megahertz.

Analog to Digital

The number of applications where analog to digital conversion makes engineering sense has increased as cost has come down and performance has gone up. ADCs are being used more and more in document scanners, camcorders and medical imaging devices. Once limited to cost insensitive military systems, direct IF sampling is now becoming more cost effective than traditional analog signal processing in some applications. "Cost is really the thing that's going to make data conversion attractive," says David Duff, Product Manager for High Speed Converter products at Analog Devices. According to Duff, both parts and labor can be reduced when digitized IF signals are demodulated and processed in digital signal processing (DSP) devices.

Sampling speed and resolution are traditionally the two most looked at figures of merit for ADCs. Both of these figures have increased, but it is still necessary to trade one for the other. According to Pat Kirk, Product Manager for High Speed Products at Burr-Brown Corporation, some speeds and bit resolutions representative of higher end devices are 16 bits at 1 Mega-sample per second (Ms/s), 14 bits at 10 Ms/s, 12 bits at 25 Ms/s and 10 bits at 100 Ms/s.

While the number of bits is certainly an indication of the resolution possible with an ADC, the effective resolution depends on several other factors. "The Effective Number of Bits (ENOB) is always less than the theoretical resolution," notes Analog's Duff, "at high analog input frequencies, ENOB is dominated less by quantization noise than by ADC AC non-linearities." An ADC's application determines what type of noise is most undesirable. According to Comlinear's Data Conversion Product Line Manager, Alan Hansford, applications that convert sampled data into the frequency domain are concerned with harmonics. For those applications that keep sampled data in the time domain, such as video applications, signal to noise ratio and differential non-linearity are most important.

Sample speed, resolution and noise are all balanced when an ADC's architecture is selected. Flash converters are the fastest, but are only used in ADCs of eight bits or less. Most high speed, high resolution ADCs today use subranging or pipelined architectures. To reduce non-linearities caused by mis-sampling an AC signal, many ADCs use sample-and-hold or track-and-hold amplifiers to give the ADC a DC value to convert.

One more concern of ADC users is power consumption. For portable applications, this concern is obvious, but even some stationary applications such as ultrasound imagers are also concerned with power consumption. "For an ultrasound machine with hundreds of channels to be digitized, it's literally a function of the current that can be pulled out of the wall socket," notes Burr-Brown's Kirk.

Digital to Analog

DACs at RF frequencies are almost exclusively used in direct digital synthesis (DDS). "DDS is an enabler, not just a more elegant way to do something, but a way to do something you couldn't do before," says Bill Woodruff, Director of Marketing at the DCSP Division of TriQuint Semiconductor. Qualcomm is currently releasing a line of DACs to go specifically with their DDS chips. "We would like to offer a complete system solution," noted Steve Anderson, a Technical Product Manager for Qualcomm, "in the past you didn't know what the spurious response of a DDS/DAC combination was until you breadboarded it." Just as with ADCs, DACs often reduce part counts because of their ability, in conjunction with a numerically controlled oscillator (NCO), to directly produce a modulated IF signal. In the case of a complex modulated signal such as 64 QPSK, the DDS approach is much less complex and much more robust than an analog approach.

While ADC designers try to strike a balance between speed and resolution, DAC designers struggle to balance speed and spurious free dynamic range. "The higher a DAC's output frequency, the more spurs you will get from the AC non-linearities in the DAC," says Burr Brown's Kirk. The same signals that are the forte of DDS, broadband and frequency agile signals, also make the design of filters to attenuate spurs exceedingly difficult. The problem of designing fast moving filters for frequency agile applications can be eliminated if the spur free dynamic range is large enough to ensure that all significant spurs will fall well outside the channel bandwidth.

Trends

ADCs and DACs are becoming less expensive, in part, because more and more of the devices are monolithic. High performance ADCs and DACs have traditionally been hybrid devices. Hybrid construction allows designers to use the optimum technology for each part of the system. Designers of monolithic converters have to compromise the performance of each sub-system to get an acceptable total system performance. The disadvantage of hybrid construction is its cost. "In high volume applications the cost of hybrids can only come down so much. But monolithically, that's when you can really reduce the cost," says Richard Mintle, Director of Marketing for Signal Processing Technology.

"The converter capacities and RF requirements are beginning to go hand-in-hand," says Don Travers, Product Marketing Manager at Analogic. It has been a long haul from concept to implementation, but RF data conversion has gained a foothold in RF designs. ***RF***

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A 20-page color brochure on Anritsu's MS2702A and MS2802A spectrum analyzers has been released by Anritsu Wiltron Sales Company. The brochure also contains information on creating a specialized measurement system using Anritsu's Personal Test Automation (PTA), and optimal system construction.
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