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engineering principles and practices

April 1995



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Cover Story
**Simulating Spread Spectrum
Systems in Software**

Plus —
**RF Active Filters
Image-Reject Mixer Tutorial**



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To receive a FREE copy of our application notes and catalog, contact Bent Hessen Schmidt at Wireless International Corp., E. 49 Midland Ave., Paramus, NJ 07652. Tel: (201) 261-8797.

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
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JMS-1LH	+10	2-500	DC-500	5.75	55	45	8.45
JMS-1MH	+13	2-500	DC-500	5.75	60	45	9.45
JMS-1H	+17	2-500	DC-500	5.90	50	50	11.45
JMS-2L	+3	800-1000	DC-200	7.0	24	20	7.45
JMS-2	+7	20-1000	DC-1000	7.0	50	47	7.45
JMS-2LH	+10	20-1000	DC-1000	6.5	48	35	9.45
JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
JMS-2H	+17	20-1000	DC-1000	7.0	50	47	12.45
JMS-2W	+7	5-1200	DC-500	6.8	60	48	7.95
JMS-11X	+7	5-1900	5-1000	6.7	35	37	4.25*

Note: *10-49 qty.

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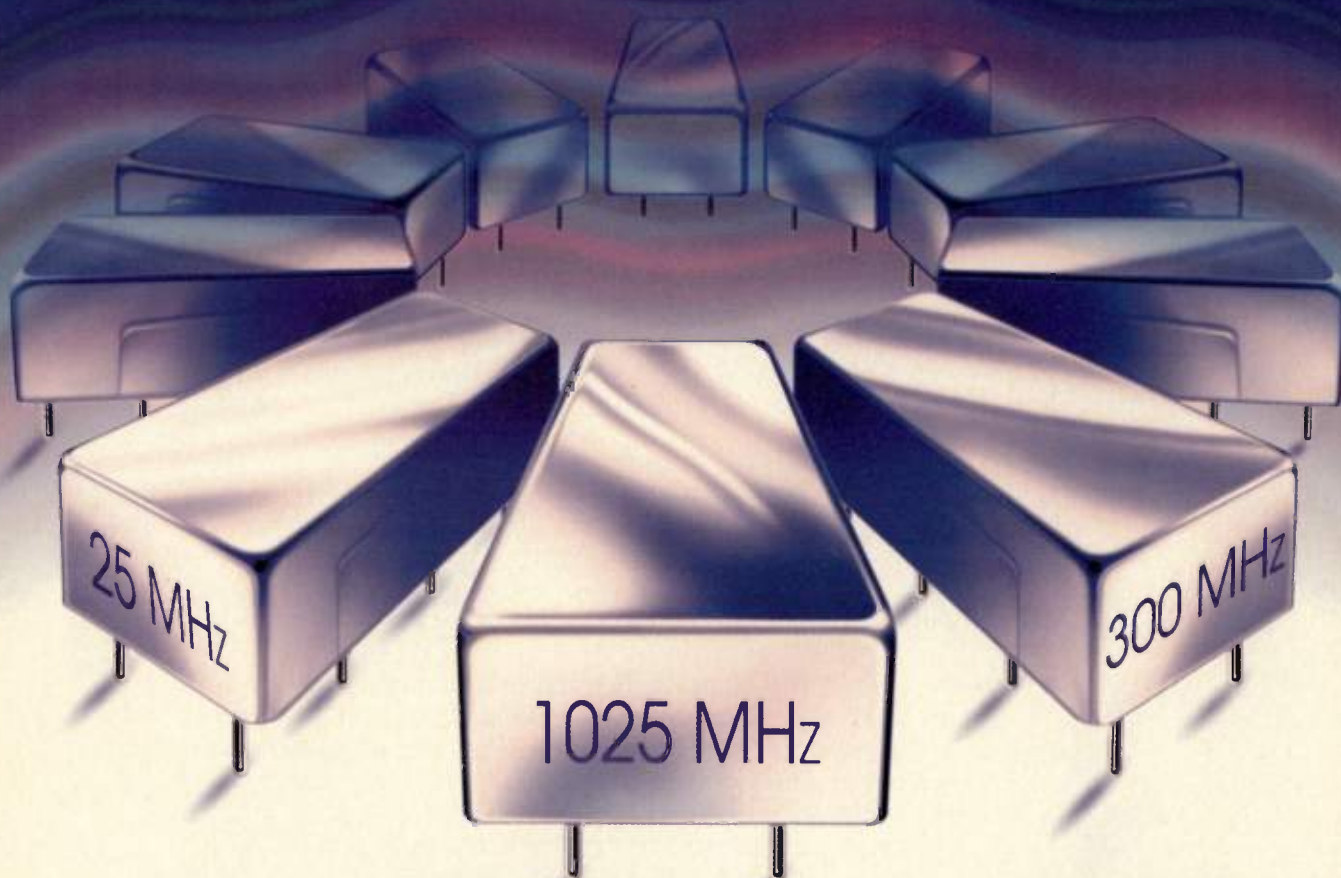
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POS-50	25-50	-110	-19	17	11.95
POS-75	37.5-75	-110	-27	17	11.95
POS-100	50-100	-107	-23	18	11.95
POS-150	75-150	-103	-23	18	11.95
POS-200	100-200	-102	-24	18	11.95
POS-300	150-280	-100	-30	18	13.95
POS-400	200-380	-98	-28	18	13.95
POS-535	300-525	-93	-26	18	13.95
POS-765	485-765	-85	-21	22	14.95
POS-1025	685-1025	-84	-23	22	16.95

Notes: Tuning voltage 1 to 16V required to cover freq. range.
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INFO/CARD 3



featured technology

24 A UHF Delay Equalizer

This article describes the effects of unequal delay on a signal and describes a circuit using a transmission line transformer to implement a delay equalizer at UHF.

—Eugene E. Mayle

30 Active Filters Using High Speed Op Amps

This article focuses on the effects of using current-feedback or voltage-feedback op amps in active filter designs. The simulated and measured results of two filters, one using a voltage feedback and one using a current feedback op amp are compared.

—Anthony D. Wang

cover story

44 Simulator Package Models a Spread Spectrum System (Part 1)

The block-level models of ELANIX' SystemView software package are used to produce functional models for a mixer, a phase detector, and a biphasic modulator.

—Stephen Kratzet

tutorial

60 Image Reject and Image Canceling Mixers

Unless some sort of suppression is used, the mixing process converts both the intended frequency and its image to an IF. Image reject and image canceling mixers use the properties of quadrature signals to suppress the image frequency.

—Louis Pandula

design awards

66 Linear Circuit Analysis Program Uses Two-Port Method

LINC (for LINEar Circuit analysis) is a circuit simulation program written for Microsoft Windows™. The program uses linear two-port analysis techniques to analyze circuits, and provides a number of graphical outputs and the ability to tune components while watching output results.

—Dale Henkes

Special Pull-Out Section

The April edition of *EMC Test & Design* is included with this issue of *RF Design*. Articles about suppression of electrical noise from DC motors and about an antenna for EMC immunity testing are included in the pull-out section.

EMC.
TEST & DESIGN



departments

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RF editorial

New Column Features Great Ideas for RF Engineers

By Gary A. Breed
Editor

The exploding demand for state-of-the-art information on RF technology and design methods has made it very hard for us to publish all the things we know you want to read about. To get one important type of information into print, we're getting ready to introduce a new addition to our magazine — something both fun and useful.

Starting this June, *RF Design* begins a new pull-out section called the "Engineer's Notebook." In this section, we will include those "great little ideas" that RF engineers are always inventing to make their jobs easier, to get a little extra performance out of a circuit, or just to "see how it works."

We encourage you and your colleagues to simply jot down an idea, sketch a diagram, and send it in! As we say in our announcement on page 71, it isn't your writing we want, it's your ideas! Anything remotely related to RF is fair game — from amplifiers to zero-crossing detectors.

We'll also track down some ideas ourselves, like circuit examples from manufacturer's application notes or past contest entries that we haven't yet had an opportunity to publish. We'll also get back to publishing book reviews on recently released titles.

The RF Design Awards Contest

To make our new Engineer's Notebook even more fun, every idea will be an entry in the next RF Design Awards contest! We're tracking down prizes right now.

Also, the contest will become an audience participation event — you, our readers, will be the judges! Every

issue's top vote-getter will win a prize, and each year, a panel of editors and past contest winners will select the best idea of the year as our Grand Prize recipient. If our prize support is as good as expected, we'll award some runner-up prizes, too.

We're excited to be able to bring RF engineers a new "grass roots" idea exchange. Even in this era of high-level computer simulation and one-chip radios, an idea sketched on the back of the latest company reorganization memo can be a powerful catalyst for a creative engineering solution.

Plan Now for RF Expo East

Our readers in the eastern U.S. have a new opportunity to attend RF Expo East — August 21-23 at the Baltimore Convention Center. We'll have a top-notch lineup of technical papers, with special emphasis on applications at the top of your priority list. We'll also have our ever-popular full-day tutorials on RF basics, oscillators, filters, and a new class on frequency synthesizers that was first presented at RF Expo West.

We're also introducing new presentations directed to marketing, sales and purchasing personnel. RF Expo is a terrific place to see a lot of suppliers in one place, so we're offering support for RF buyers and sellers, updating them on trends in the technology, component developments and future market possibilities.

I hope to see you at the beautiful Baltimore Inner Harbor, where RF Expo East is this year's number one place for RF business and technology!



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*727LC	10W CW	.006-1000 MHz	44dB	\$ 7,950
713FC	15W CW	20-1000 MHz	42dB	\$ 5,680
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
714FC	30W CW	20-1000 MHz	45dB	\$ 9,350
250LC	50W CW	.01-225 MHz	47dB	\$ 5,550
715FC	50W CW	200-1000 MHz	47dB	\$ 14,990
707FC	50W CW	400-1000 MHz	50dB	\$ 10,990
716FC	50W CW	20-1000MHz	47dB	\$ 17,950
*747LC	50W CW	.01-1000 MHz	47dB	\$ 18,550
116FC	100W CW	.01-225 MHz	50dB	\$ 9,500
709FC	100W CW	500-1000 MHz	50dB	\$ 16,990
717FC	100W CW	200-1000 MHz	50dB	\$ 19,500
718FC	100W CW	20-1000 MHz	50dB	\$ 29,800
7100LC	100W CW	80-1000 MHz	50dB	\$ 19,500
*757LC	100W CW	.01-1000 MHz	50dB	\$ 29,950
122FC	250W CW	.01-225 MHz	55dB	\$ 19,950
723FC	300W CW	500-1000 MHz	55dB	\$ 29,995
LA500V	500W CW	10-100 MHz	56dB	\$ 12,900
LA500UF	500W CW	100-500 MHz	57dB	\$ 46,000
LA500G	500W CW	500-1000 MHz	57dB	\$ 55,000
LA1000V	1000W CW	10-100 MHz	60dB	\$ 22,500
LA1000UF	1000W CW	100-500 MHz	60dB	\$ 75,000
LA1000G	1000W CW	500-1000 MHz	60dB	\$ 99,000
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RF calendar

April

- 23-26 IEEE Instrumentation/Measurement Technology Conference**
Waltham, MA
Information: Robert Myers, 3685 Motor Ave., Suite 240, Los Angeles, CA 90034. Tel: (310) 287-1463. Fax: (310) 287-1851, or Dan Sheingold, Analog Devices, P.O. Box 280, Norwood, MA 02062. Tel: (617) 461-3294. Fax: (617) 329-1241.
- 25-27 International Wireless Communications Expo 95**
Las Vegas, NV
Information: IWCE 95, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828-0420. Fax: (404) 618-0441.
- 25-28 Internecon/Semiconductor Shanghai 95**
Shanghai, China
Information: Joe Nemchek, International Sales Force, Reed Exhibition Companies, 383 Main Avenue, Norwalk, CT 06851. Tel: (203) 840-5398. Fax: (203) 840-9398.

May

- 14-19 IEEE/MTT-S Microwave Symposium**
Orlando, FL
Information: 1995 IEEE Symposium, c/o Horizon House Publications, 685 Canton Street, Norwood, MA 02062. Fax: (617) 762-9230.
- 21-24 45th Electronic Components and Technology Conference**
Las Vegas, NV
Information: Jim Bruorton, Publicity Chairman, c/o KEMET Electronics Corporation, P.O. Box 5928, Greenville, SC 29606. Tel: (803) 963-6621. Fax: (803) 963-6521.
- 31-2 1995 Virginia Tech Symposium on Wireless Personal Communications**
Blacksburg, VA
Information: Jenny Frank, Administrator, Mobile and Portable Radio Research Group. Tel: (703) 231-2958.

June

- 1-2 ENDIEL - The Portuguese Electric and Electronic Products Exhibition**
Lisbon, Portugal
Information: Silicon Electronica E Telematica, Edificio Pascoal de Melo, Rua Pascoal de Melo, N. 3, 1100 Lisboa, Portugal. Tel: 8151234. Fax: 8130796.
- 13-15 Nepcon East**
Boston, MA
Information: Reed Exhibition Companies, 383 Main Avenue, Norwalk, CT 06851. Tel: (203) 840-5398. Fax: (203) 840-9398.
- 21-23 Electro/International 1995**
Boston, MA
Information: Miller Freeman, Kathryn Piersall, 13/6D Noel Road, Suite 500, Dallas, TX 75240. Tel: (214) 419-7969. Fax: (214) 419-7915.

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MAX3M050055	1.67-1.83	5.0-5.5
MAX3M055060	1.83-2.0	5.5-6.0
MAX3M060065	2.0-2.16	6.0-6.5
MAX3M065070	2.16-2.33	6.5-7.0
MAX3M070075	2.33-2.5	7.0-7.5
MAX3M075080	2.5-2.66	7.5-8.0
MAX3M080085	2.66-2.83	8.0-8.5
MAX3M043052	1.43-1.73	4.3-5.2
MAX3M047056	1.56-1.86	4.7-5.6
MAX3M063074	2.1-2.46	6.3-7.4
MAX3M070083	2.3-2.76	7.0-8.3

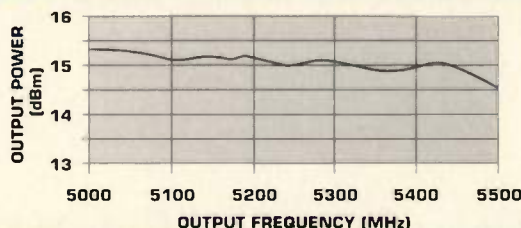
FREQUENCY QUADRUPLERS

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MAX4M055060	1.375-1.5	5.5-6.0
MAX4M060065	1.5-1.625	6.0-6.5
MAX4M065070	1.625-1.75	6.5-7.0
MAX4M070075	1.75-1.875	7.0-7.5

FREQUENCY QUINTUPLERS

MAX5M085090	1.7-1.8	8.5-9.0
MAX5M090095	1.8-1.9	9.0-9.5
MAX5M095105	1.9-2.1	9.5-10.5
MAX5M105115	2.1-2.3	10.5-11.5
MAX5M115125	2.3-2.5	11.5-12.5
MAX5M125135	2.5-2.7	12.5-13.5
MAX5M135145	2.7-2.9	13.5-14.5
MAX5M114127	2.28-2.56	11.4-12.8
MAX5M127142	2.54-2.84	12.7-14.2

MAX3M050055 - OUTPUT POWER VS. FREQUENCY



COMMON ELECTRICAL SPECIFICATIONS

Gain.....	3 dB typical
Input drive	8-12 dBm
Output level.....	11-15 dBm
Spurious rejection.....	60 dBc minimum
Input harmonic rejection	60 dBc minimum
Harmonics of output	15 dBc typical
Output power flatness	±1dB typical
VSWR, in/out	2 1/1.5 1 typical

ENVIRONMENTAL SPECIFICATIONS

Operating temperature	-30° to +70°C
Storage temperature.....	-30° to +85°C
Humidity	95% relative humidity, noncondensing
Vibration.....	7 Gs RMS, 50-5000 CPS, per MIL-STD-810B, Method 514, Procedure 5

ADDITIONAL INPUT POWER OPTIONS

Option	Input Drive (dBm)	Output Power (dBm)
J	3-8	6-11
H	12-16	12-16

In addition to the frequency multipliers listed here, MITEQ offers a line of passive and active frequency doublers and higher order multipliers to 40 GHz. For additional information please call Dave Krautheimer at extension 187.



WARRANTY

RF courses

High-Frequency Analog Circuit Design for Communication Systems

June 12-15, 1995, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson.
Tel: (46) 122-175-70. Fax: (46) 122-143-47.

Antennas and Antenna Systems: Practical Design, Implementation, and Testing

April 24-27, 1995, Washington, DC
Modern Digital Modulation Techniques
May 1-5, 1995, Washington, DC

Modern Receiver Design

June 12-16, 1995, San Diego, CA
Information: The George Washington University,
Continuing Engineering Education, Academic Center,
Room T-308, 801 22nd Street, N.W., Washington, DC 20052.
Tel: (202) 994-6106 or (800) 424-9773. Fax: (202) 872-0645.

Applied RF Techniques II

May 1-5, 1995, Middletown, NJ

Applied RF Techniques

May 8-12, 1995, Los Altos, CA
Fundamentals of Communication Technologies

May 10-12, 1995, Los Altos, CA

Wireless RF System Design

May 15-19, 1995, Los Altos, CA

RF/MW Measurement Techniques I

June 12-16, 1995, Cambridge, UK

Applied RF Techniques I

June 12-16, 1995, Cambridge UK

RF Component Modeling

June 19-22, 1995, Cambridge UK
Information: Besser Associates, 4600 El Camino Real,
Suite 210, Los Altos, CA 94022. Tel: (415) 949-3300.
Fax: (415) 949-4400.

Wireless Digital Communications

May 8-12, 1995, Tempe, AZ
Information: Arizona State University, Center for
Professional Development, Box 877506, Tempe, AZ
85287-7506. Tel: (602) 965-1740. Fax: (602) 965-8653.

RF IC Design for Wireless Communication Systems

April 24-28, 1995, Santa Clara, CA
Information: MEAD Microelectronics, Inc., 7100 NW
Grandview Drive, Corvallis, OR 97330. Tel: (503) 758-0828.
Fax: (503) 752-1405.

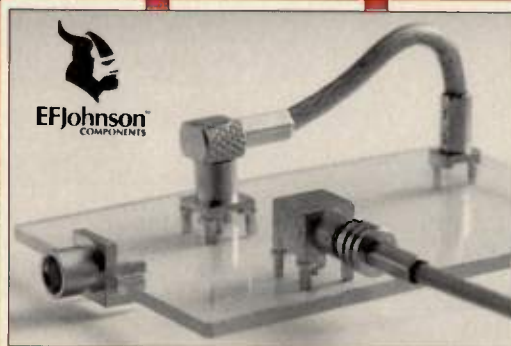
Real-Time Digital Signal Processing

May 16-18, 1995, Kansas City, KS
Design of High-Performance Wireless Communication Systems

May 16-18, 1995, Kansas City, KS
Information: The University of Kansas, Attn: Lorene
Damewood, Continuing Education Building, Lawrence, KS
66045-2607. Tel: (913) 864-3284. Fax: (913) 864-5074.

We're Into Real Estate!

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MCX Connectors by E.F. Johnson Company

With the shortage of real estate - why give up valuable board space to larger RF connectors? Our MCX is 30% smaller than SMB connectors and an excellent choice for high density packaging requirements. With a rugged snap-on mating design, these crimpable contact connectors allow for fast and dependable crimp-crimp assembly. Our MCXs employ .020" standoffs for improved board cleaning and are available with .100" leg spacing. The original E.F. Johnson Company end launch connector is also part of the MCX family.

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E.F. Johnson Company, 299 Johnson Ave., Waseca, MN 56093 Fax: (507) 835-6287

INFO/CARD 11

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- Low phase noise models
- S/C/X/Ku-Band modules
- 70 MHz or 140 MHz IF
- INTELSAT and EUTELSAT standards
- Low microphonic design
- 3.5 inches high, 19-inch rack

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elisra electronic systems ltd.

48 Mivtza Kadesh Street, Bene Beraq 51203, Israel
Tel: 972-3-7545655/5639, Fax: 972-3-7545299

INFO/CARD 12

ARRL Radio Designer

Heavyweight RF CAD features without a heavyweight price

If you've done any hobby computer-aided design, you've probably used CAD mainly for designing and enhancing your antenna system. Now with your PC and *ARRL Radio Designer* you'll be able to build, test and evaluate audio and radio circuits without warming up your soldering iron, running out of parts, or investing in a benchful of expensive test equipment.

Here's what you and *ARRL Radio Designer* will accomplish quickly and easily:

- Model passive and small-signal linear circuits from audio to RF.
- Do "what if" circuit modifications without tedious cut-and-try experimentation and measurement.
- Predict and analyze the performance of linear, small-signal active and passive dc, audio and RF circuitry, including amplifiers, filters, matching networks and power splitters and combiners.
- Enter circuits in netlist form using the text-based Circuit Editor.
- Optimize circuit performance to meet goals you specify.
- Display the signal level at any point in a simulated circuit with Voltage Probe.
- Simulate component value variations due to temperature and tolerance with Monte Carlo statistical analysis.
- Simulate circuit performance in response to a steady-state time-domain signal using impulse, step, pulsed carrier or user-defined stimuli.
- Synthesize matching networks with Circles, an interactive Smith® Chart utility.
- Use manufacturer-supplied noise and S-parameter device data for highly realistic circuit simulations.

ARRL Radio Designer 1.0's 28 Circuit elements include

- Active devices (bipolar and field-effect transistors, operational amplifier)
- "Black box" elements (two-terminal impedance specified by resistance and reactance; two-terminal one port specified by admittance, impedance or reflection coefficient; three-terminal two-port specified by admittance, impedance or S parameters)
- Controlled sources (current-controlled current, current-controlled voltage, voltage-controlled current, voltage-controlled voltage)
- Lossy capacitors, inductors, coaxial cables and transmission lines
- Transformers (ideal two- and three-winding types, specifiable in terms of turns ratio or impedance)

Technical Specifications for ARRL Radio Designer 1.0 Analysis:

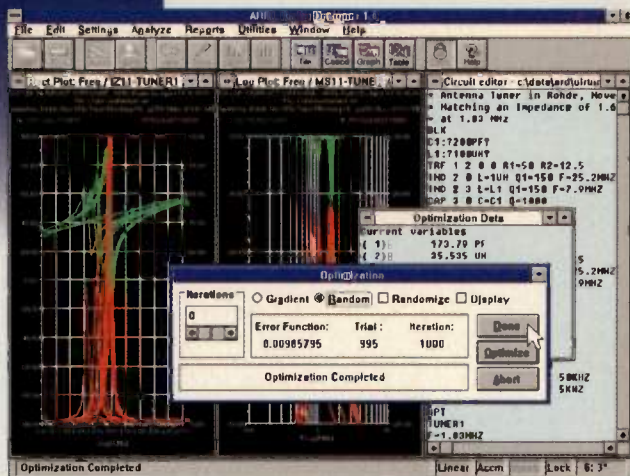
Maximum number of nodes per circuit block*	250
Range of node-number values	0 through 999
Maximum number of ports per circuit block	30
Maximum number of frequency steps	512
Maximum number of statistical histograms	20

* The number of circuit blocks allowed depends on their complexity; generally, 50 or more circuit blocks may be defined in a single *ARRL Radio Designer* netlist.

ARRL Radio Designer is a derivative of Super-Compact®, an industry-standard linear circuit simulator by Compact Software of Paterson, New Jersey.

ARRL 225 Main St. • Newington, CT 06111-1494
Phone (203) 666-1541

INFO/CARD 13



ARRL Radio Designer simulates and reports

- S, Y, Z, group delay and voltage probe parameters for n -port networks;
- Chain (ABCD), hybrid (H), inverse hybrid (G), gain, voltage gain, and stability parameters for two-port networks;
- Magnitude of reflection coefficient, phase of reflection coefficient, VSWR and return loss parameters for one-port networks;
- Gain, gain matching and noise parameters; and
- Complex S, Y, Z, H, G, chain (A), gain matching, noise matching and voltage probe parameters.

ARRL Radio Designer reports can be—

- Rectangular and polar graphs—onscreen and via any Windows™ compatible printer, in the colors, fonts and line weights you specify
- Tables—onscreen, in file form and via any Windows™ compatible printer

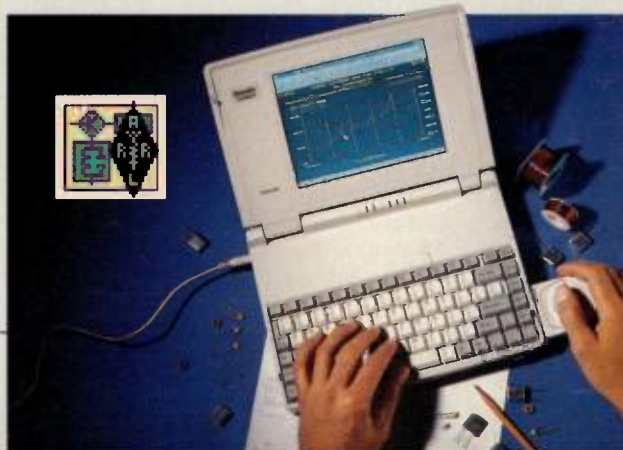
Minimum system requirements for ARRL Radio Designer 1.0:

- 386, 486 or Pentium™ IBM PC or 100% compatible (math coprocessor strongly recommended)
- 8 Mbytes of RAM
- Microsoft Windows™ 3.1 or higher
- 3.5-inch, high-density floppy drive
- hard disk with at least 5 Mbytes of free space
- Mouse or equivalent pointing device

We re pleased to offer

ARRL Radio Designer as your doorway into a new phase of ham radio's growing computer tradition. The package includes two 3.5" diskettes, circuit examples and a comprehensive reference manual.

ARRL Order No. 4882 Retail \$150 (add \$5 for UPS delivery).



Report Lists Top 10 Technologies

Battelle researchers have produced a list of the top ten strategic technologies for 2005. "One of the most striking overall advances will be in the area of miniaturization," said Stephen Millett, Managing Principal of the Battelle Technology Management Group.

The choices for the ten strategic technologies were based on three criteria. They must: provide benefit to the end user, i.e. will someone buy it?; enjoy a protected and sustainable competitive advantage in such areas as quality, uniqueness, or price; and support business goals.

The top ten strategic technologies, in order of importance, are:

1. Mapping of the human genome for genetic-based personal identifica-

tion and diagnosis.

2. Super materials.

3. Compact, long-lasting and highly portable energy sources, including fuel cells and batteries.

4. Digital high definition television.

5. Miniaturization of electronics for personal use.

6. Cost-effective systems that integrate power, sensors, and controls.

7. Anti-aging products and services.

8. Medical treatments, with highly accurate sensors and problem locators, and drug delivery systems that will be highly specific to precisely targeted parts of the body.

9. Hybrid fuel vehicles.

10. Edutainment - educational games and computerized simulations.

NIST Upgrades FMAS

The National Institute of Standards and Technology's Frequency Measurement and Analysis Service (FMAS) has been upgraded in resolution, accuracy and capabilities. The FMAS now offers better than 40-picosecond single-shot resolution; one part in ten trillion (1×10^{-13}) accuracy over 24 hours; and it now computes short term stability using the Allan variance.

Subscribers to the FMAS are provided with all necessary equipment, including a 486-class computer and monitor, GPS receiver and antenna, time interval counter, modem, uninterruptible power supply, tape backup, and printer. Training at the NIST Boulder, CO site and complete technical support is also provided.

Thermo Voltek Acquires Kalmus Engineering

Thermo Voltek Corporation has announced that its KeyTek Instrument Division has acquired substantially all of the assets of Kalmus Engineering Incorporated and its subsidiary, R.F. Power Labs, Inc. Kalmus is a manufacturer of radio frequency power amplifiers and systems used in determining the immunity of other electronic products to radiated or conducted radio frequency interference, and in medical imaging and telecommunications. Thermo Voltek Corp. designs, manufacturers, and markets instruments that test electronic systems and components for immunity to electromagnetic

interference and provides related distribution and consulting services.

Call for Papers

A call for papers has been released for the Sixth International Conference on Signal Processing Applications and Technology, ICSPAT '95. The conference will be held October 24-26, 1995 in Boston, MA. Application areas for papers include aerospace, audio and speech, comm. and telephony, consumer products, data acquisition, DSP hardware, DSP software, geophysics, image processing, industrial applications, instrumentation and testing, medical electronics, multimedia, neural networks, parallel processing, radar, radio, SATCOM and NAV, real-time O/S, robotics, speech processing, underwater/SOAR, and VLSI architectures. Mail, Fax or E-Mail 400 word abstracts by April 15, 1995 to ICSPAT Staff, DSP Associates, 49 River Street, Waltham, MA 02154. Tel: (617) 891-6000. Fax: (617) 899-4449. E-Mail: icspat@dspnet.com.

Contest Winners

Paul Peterzell of Qualcomm and Guy Tant of The County of San Diego were the winners of the airline tickets that were awarded for the *RF Design* video contest held during the RF Expo West — EMC/ESD International Conference, January 29-February 1, 1995 in San Diego. Special thanks to Jennifer Green of JFW Industries who drew the winners. And, thanks to all who attended the show and participated.

Contracts

Motorola to Build Phone System in Columbia — Motorola's Pan American Market Division has signed two separate contracts to deploy three fixed wireless phone systems in Columbia. The installation of Motorola's Wireless Local Loop (WILL®) product will bring phone service for the first time to many residents in remote areas and complement the country's existing telephone systems in the cities of Bogota, Cali and Subachoque.

Milcom Awarded Contract for First CDMA System in Korea — Milcom International, Inc. has been awarded multi-million dollar contracts from two major electronics firms for Linear Amplifiers for the world's first CDMA system in Seoul, Korea. Shipments of the linear MCA8000-250 will commence in the first quarter of 1995.

Centurion and Ericsson Sign Agreement — Centurion International, Inc. signed a four-year \$100 million agreement with the cellular phones division of Ericsson, Inc. Centurion will provide \$100 million in nickel cadmium and nickel metal hydride batteries to meet all of Ericsson's world wide battery requirements.

\$2.5 Million for Wireless Communications Research — Two grants totaling \$2.5 million to improve wireless communication technology were awarded to Virginia Tech's Mobile and Portable Radio Research Group (MPRG). The Advanced Research Projects Agency has awarded MPRG a \$1.7 million contract over the next three years. And the Office of National Drug Control Policy is funding another \$800,000 for the MPRG to develop a new surveillance and tracking software for installing wireless communication systems in law enforcement and military applications.

GTE and ATG Sign Purchase Contract — Allen Telecom Group has reached a 3-year, non-exclusive open purchase agreement with GTE Communication Systems Corporation. The multimillion dollar contract provides an open procurement avenue for all of GTE's domestic and international business units whenever and wherever the need for wireless communications equipment should arise.

Richardson Electronics, Ltd.
named Worldwide Distributor
for Stanford Microdevices

Stanford Microdevices has signed a distribution agreement with Richardson Electronics for worldwide coverage. Richardson will carry all of Stanford Microdevice's standard products including both discrete and MMIC product lines.

Richardson Electronics is an international distributor of power semiconductors used primarily in industrial and telecommunications applications.

Stanford specializes in discrete and MMIC amplifiers with output power in the half-watt to 10 watt range.

4 Watt Amp



Designed for operation in wireless systems operating in the 1500 to 2500 MHz frequency range, this amplifier has 26 dB of gain with 36 dBm of output power at P1dB.

INFO/CARD 118

1 Watt, 2 Watt and 4 Watt Amplifier



**Stanford's
#1 MMIC!**

These high-performance GaAs MMIC devices are housed in copper-tungsten packages for efficient heat-transfer and are 100% tested at +85°C to ensure reliable field performance.

	Frequency	Gain*	P1dB*
SMM-210	1.5-2.5 GHz	25 dB	30 dBm
SMM-280-2	1.5-2.5 GHz	25 dB	33 dBm
SMM-280-4	1.5-2.5 GHz	25 dB	36 dBm

*Minimum at 2 GHz.

INFO/CARD 14

1/2 Watt Discrete FETs

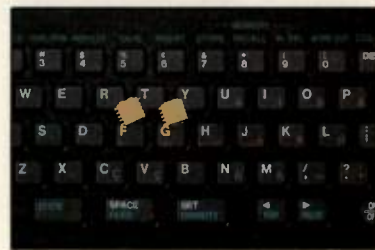


COMING SOON!

A new line of power MESFETs offering 1/2 Watt to 5 Watts with high efficiency. Call 1-800-RF-POWER for technical information.

INFO/CARD 117

Broadband LNAs



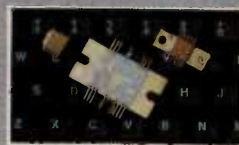
Designed for operation in wireless applications operating in the 100-4000 MHz frequency range, these LNAs are also an ideal choice for applications requiring flat gain, low noise figure and low VSWR over a very wide band.

	Frequency	Gain	P1dB
SMM-008	0.1-2.5 GHz	18 dB	15 dBm
SMM-010	0.1-4.0 GHz	18 dB	15 dBm

INFO/CARD 116

**Stanford
Microdevices**

MMIC POWER AMPS



**NEW
PRODUCTS**

Part Number	Frequency (GHz)	Gain (dB)	P1dB (dBm)	TOIP (dBm)	Min Supply Voltage (V)	Package
SMM-208	1.5-2.5	25	+28	+37	+3.0	plastic so-8
SMM-210	1.5-2.5	25	+30	+39	+5.0	10-pin ceramic
SMM-280-2	1.5-2.5	25	+33	+42	+7.0	0.6 x 1.0" flange
SMM-280-4	1.5-2.5	25	+36	+45	+7.0	0.6 x 1.0" flange
SMM-610	5.9-6.4	27	+33	+42	+5.0	10-pin ceramic
SM-680-2	5.9-6.4	27	+36	+45	+7.0	0.6 x 1.0" flange
SMM-1820-1	6.0-18	11	+27	+36	+8.0	0.6 x 1.0" flange
SMM-1820-2	6.0-18	11	+30	+39	+8.0	0.6 x 1.0" flange

BROADBAND MMIC AMPS

Part Number	Frequency (GHz)	Gain (dB)	Noise Figure (dB)	Min Supply Voltage (V)	P1dB (dBm)	Package
SMM-008	0.1-2.5	18	2.4	+5.0	+15	plastic so-8
SMM-010	0.1-4.0	18	2.2	+5.0	+15	10-pin ceramic
SMM-108	0.5-2.0	18	3.5	+5.0	+12	plastic so-8
SMM-110	0.5-3.0	19	3.2	+5.0	+12	10-pin ceramic
SMM-808	2.0-7.0	13	5.0	+12	+17	plastic so-8
SMM-810-1	2.0-8.0	14	5.0	+12	+17	10-pin ceramic
SMM-810-2	2.0-8.0	27	5.0	+12	+17	10-pin ceramic
SMM-1810	6.0-18	12	5.0	+5.0	+12	10-pin ceramic
SMM-2010	2.0-18	7	5.5	+5.0	+20	10-pin ceramic

DISCRETE PHEMT'S

Part Number	Ga @ 1 GHz (dB)	Nf @ 1 GHz (dB)	Ga @ 2 GHz (dB)	Nf @ 2 GHz (dB)	P1dB (dBm)	Package
SPF-284	18	0.5	16	0.8	+10	85 mil plastic
SPF-484	18	0.3	16	0.5	+10	85 mil plastic

Part Number	Ga @ 4 GHz (dB)	Nf @ 4 GHz (dB)	Ga @ 12 GHz (dB)	Nf @ 12 GHz (dB)	P1dB (dBm)	Package
SPF-684	15	0.7	9	1.5	+10	85 mil plastic
SPF-884	15	0.5	9	1.2	+10	85 mil plastic
SPF-1076	16	0.4	11	1.1	+15	70 mil ceramic
SPF-1276	16	0.4	11	1.0	+10	70 mil ceramic
SPF-1376	15	0.3	10	0.8	+9	70 mil ceramic
SPF-1476	15	0.3	10	0.7	+9	70 mil ceramic
SPF-1576	15	0.2	10	0.6	+9	70 mil ceramic

INFO/CARD 119

Stanford Microdevices Distributed by

**Richardson
Electronics, Ltd.**

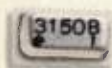
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455KHz/10MHz Filter
Offers excellent sensitivity, optimum stop band attenuation and high stability for radio applications.



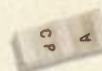
SMD Low Profile Trim Pot — SMT POZ3 Series
Our ultra-low profile component for radio tuners — just 1.8mm high with a 3mm footprint.



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Highly miniaturized, very lightweight, exceptionally stable. For use in RKE applications.



GYROSTAR
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SMD Resonator
Very small and lightweight with high stability. Available in frequencies up to 33MHz.



A few of the important compon

What goes into an ideal weekend drive? A sunny day, a winding country road, a sizable number of Murata components. In fact, the average automobile could easily contain more than 300 Murata electronic components — from microwave VCOs and ceramic resonators to vibrating gyroscopes and chip capacitors.

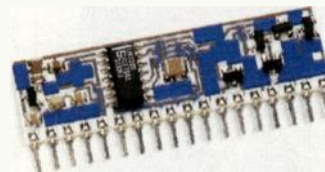
Our commitment to research (we spend more than 7% of sales on R&D — nearly twice the average in our industry) has made us the world leader in passive component technology. That technology optimizes every step of our vertically integrated manufacturing process, from raw ceramic material production

**SMT Ferrite Chips —
BLM Series**

Effectively prevents oscillation in HF circuits, to several hundreds of MHz. Sizes down to 0603.

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Ultra-miniature sizes — all the way down to 0402. Temperature ranges from -55°C to +150°C.



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For multiplexing/node wiring system, power windows, ECU and airbags, backed by Murata's 25 years of thick film technology.



Compact VCOs

Small size and low power consumption. Frequency range from 400MHz-2.5GHz. Stability of ± 2 MHz from -35°C to +80°C.

Monolithic LC Chip Filter

A cost-effective alternative to SAW filters for cellular and GPS applications. As small as 4.5 x 3.2 x 2.0mm.

its of a pleasant Sunday drive.

to the design and fabrication of complex modules like hybrid ICs for sophisticated automotive systems. You see, we not only want to be an important component of your Sunday drive. We also want to be a vital part of your designs from Monday through Friday.



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RF news *continued*

Business Briefs

Savi Technology and Digital Wireless Corp team up on Inventory Management System — Savi Technology Inc. and Digital Wireless Corporation announced that Savi has incorporated Digital Wireless' spread spectrum radio technology into Savi's Asset and Transportation Management systems. Savi uses Digital Wireless' WIT915 Recombinant Spread Spectrum™ transceiver to create a wireless information network for tracking and managing critical inventory.

Kay Elemetrics Moves — Kay Elemetrics Corp. has moved to 2 Bridgewater Lane, Lincoln Park, NJ 07035-1488.

ITT Cannon and Elcan Technologies Sign Distribution Deal — ITT Cannon Datacom has signed a nationwide reseller agreement with Elcan Technologies. The agreement will allow Elcan to provide sales, installation and technical support services in critical, high-growth markets for ISCS, ITT Cannon's Structured Cabling System based on Screened Twisted Pair cabling technology.

Scientific-Atlanta Selected by Korea Telecom — Scientific-Atlanta has been selected as the preferred provider of satellite communications equipment for Korea Telecom's new cable television distribution network. The Korea Telecom network will provide cable television programming via satellite to cable headends throughout the Republic of Korea.

AEG and Andrew Join Forces — AEG Transportation Systems, Inc. and Andrew Corporation, Communications Systems Group, have teamed up to develop a communications based advanced automatic train control system (Flexiblok™). When completed, the spread spectrum radio based system could replace the traditional track circuit approach which employs electrical track shorting for vehicle detection, positioning and speed code transmission.

Glenayre to Merge with Western Multiplex — Glenayre Technologies, Inc. and Western Multiplex Corporation have signed a definitive agreement to merge the two companies. Western Multiplex design, manufacturers and sells digital and analog point-to-point microwave radios for a variety of markets. They will operate as the Western Multiplex Division of Glenayre and will continue to market products under their brand name and trademarks.

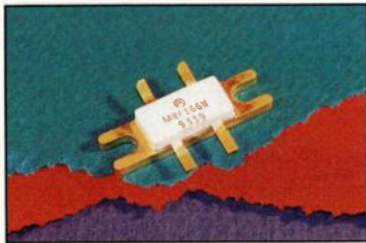
Epitronics Announces Contract Award — Epitronics Corporation announces a recent Phase II SBIR (Defense Small Business Innovation Research Program) contract award sponsored by Advanced Research Projects Agency. The award to continue investigation of Indium Phosphide based device structures for microwave and millimeter wave monolithic integrated circuits centers around developing and characterizing InP based HEMT and PHEMT structures.

Arnold Announces Plant Upgrade — Arnold Engineering Corporation has announced the completion of a \$5 million upgrade of the MPP Core facility. The major benefits of the upgrade include: increased manufacturing capacity, reduced lead times, and a more uniform product with superior magnetic properties.

SGS-Thompson Opens New Design Center — SGS-Thompson Microelectronics has set up its own building to house its design center in Noida, India. The 5000 sq. m. building will carry out a variety of projects, providing quality solutions to SSG customers worldwide.

Racal Drops Prices — Increased volumes enable Racal Instruments to announce a considerable price reduction on high-power 13-slot VXIbus chassis and three of the best selling switches.

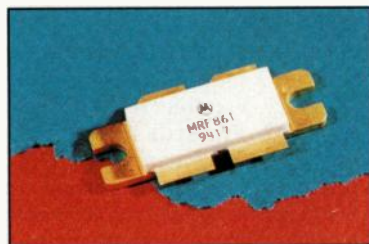
**MRF166W - Class A, 40 W, 500 MHz,
Broadband Power MOSFET**



Motorola's rugged 40 watt RF Power MOSFET is designed primarily for wideband large-signal output and driver stages to 500 MHz. The MRF166W's excellent thermal stability makes it ideal for Class A operation. Its push-pull configuration reduces even numbered harmonics. Other features include a 13 dB minimum gain and 50% efficiency at 40 watts.

Richardson Electronics, Ltd.
40W267 Keslinger Road, LaFox, IL 60147.
1-800-348-5580.
INFO/CARD 113

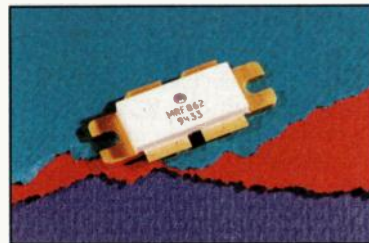
**MRF861 - Class A, 800-960 MHz,
27 W (CW) RF Power Transistor**



Motorola's 27 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include a 9.5 dB minimum gain, an output capacitance of 45 pF at 24 Vdc, a minimum ITO of +53.5 dBm and typical noise figure of 6.5 dB. The MRF861 will withstand RF input overdrive of 8 W CW. It is in a push-pull flange package (case 375A).

Richardson Electronics, Ltd.
40W267 Keslinger Road, LaFox, IL 60147
1-800-348-5580.
INFO/CARD 114

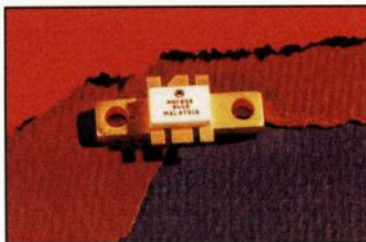
**MRF862 - Class A, 800-960 MHz,
36 W (CW) RF Power Transistor**



Motorola's 36 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include a 9 dB minimum gain, an output capacitance of 75 pF at 24 Vdc, a minimum ITO of +55 dBm and typical noise figure of 6.5 dB. The MRF862 will withstand RF input overdrive of 13.6 W CW. It is in a push-pull flange package (case 375A).

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**MRF858 - Class A, 800-960 MHz,
3.6 W (CW) RF Power Transistor**



Motorola's 3.6 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include an 11 dB minimum gain, an output capacitance of 6.5 pF (typ) at 24 Vdc, a minimum ITO of +44.5 dBm and typical noise figure of 6 dB. The MRF858 will withstand RF input overdrive of 0.85 W CW. It is in an SOE flange package (case 319).

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Keslinger Road, LaFox, IL 60147
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**MRF860 - Class A, 800-960 MHz,
13.7 W (CW) RF Power Transistor**



Motorola's 13.7 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include an 11 dB minimum gain, an output capacitance of 21 pF (typ) at 24 Vdc, a minimum ITO of +51.5 dBm and typical noise figure of 6.5 dB. The MRF860 will withstand RF input overdrive of 3.25 W CW. It is in a push-pull flange package (case 395B).

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Antennas See Bigger Markets, Better Design Tools

By Andy Kellett
Technical Editor

Antennas are the quintessential RF device. Wires can transmit all kinds of signals, and ICs are built that flip bits, but when you see an antenna, you know you are dealing with an RF system. As more people use wireless devices, manufacturers have to produce antennas more economically and have to produce antennas with consumer's requirements and prejudices in mind. At the same time, the demand for antennas for larger scale communications such as broadcasting and satellite communications are in as great a demand as ever.

Antenna Markets

The most active antenna markets are of course those that serve the most active RF markets. Many antenna makers are looking forward to the boom that should come with the installation of PCS systems across the country. At the same time older markets supply at least a steady stream of orders for antenna makers to fill.

The cellular market continues to demand more antennas as more cells are added and service providers expand their digital cellular systems. Broadcasters are steadily replacing VHF television and radio antennas. UHF television is experiencing a boom, particularly Fox network affiliates, and those stations are upgrading their antennas to improve coverage. The same expansion of UHF has caused a slight elevation in transportable satellite stations as UHF stations expand their programming to include news. On a smaller scale, antennas for intra-office communications via wireless LANS and wireless PBXs are currently a small market, but promise to grow.

Antenna Specifications

The specifications that customers are most concerned with depend on the type of antenna being considered.

In the market for cellular and PCS basestation antennas, customers are most concerned with cost, of course, but they are also concerned with radiation pattern control and gain says Dr. Thomas Charelton, Head of the Antenna Section for Andrew Corporation. For subscriber units, the most important concern is reliability, which again means concern for gain and a uniform gain pattern. While the end user of subscriber unit antennas may not be savvy to measurements in dBi, "They are very savvy about dropped calls," says Dale W. Horn, Vice President of Engineering for the Antenna Specialists Co. division of the Allen Telecom Group.

Antenna radiation patterns are critically important to broadcast antenna users, says Andrew's Charelton, because broadcasters are paid to reach the people in a specific geographic region. Another concern, of almost equal concern, is the transmitting antenna's power handling capability. "With the powers being used in these antennas, they can self destruct when something goes wrong," says Charelton. Another concern is intermodulation, particularly for broadcasting antennas overseas, where often several stations share the same antenna.

For antennas designed to uplink signals to C- and Ku-band satellites, side-lobe control is very important to reduce the amount of signal inadvertently transmitted to satellites adjacent to the intended satellite.

Antenna Design

Antenna designers have a number of software tools available to them to help them predict the behavior of almost any collection of conductors and dielectrics. NEC and its faster, but restricted version, MININEC, are in widespread use and are the core of some commercial antenna design programs. While algorithms for solving

electromagnetics problems have improved, the hardware on which those algorithms are implemented has been the main reason designers can simulate more complex antennas in less time than they have in the past.

Simulation can get a designer closer to a final design, but rarely does a computer spit out a manufacturable design on the first try. For that reason, antenna test ranges are still important to the industry. Antenna Specialists' Horn notes that his company has taken great pains to equip their antenna test range with the best instrumentation available.

Antennas and System Design

While antenna engineers have powerful tools with which to design antennas to the most exacting standards, many customers do not take advantage of that fact. The antenna is often one of the last considerations of system designers," says Dennis McGivern, Director of Marketing for AntennaCo. Andrew's Charelton sees much the same thing, "[Antennas] are left to the last minute because people think they are easy." McGivern notes that many of the specifications for the electronic components downstream from the antenna can be less strict if the proper antenna is used. Tradeoffs between mechanical and electrical performance could also be reduced if mechanical and electrical engineers on both sides of the antenna supplier/customer relationship are involved in the early design stages of an antenna says Mark Cockson, Marketing and Strategic Business Development Manager for Centurion International.

The function of an antenna cannot be shrunk and integrated into a nondescript black box the way other RF functions can. However, antenna designers are better able to make sure antennas work no matter what environment they are used in.

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A UHF Delay Equalizer

By Eugene E. Mayle
R.L. Drake Co.

High performance and efficient use of RF requires that energy be band limited. Band limiting at radio frequencies takes the form of bandpass filters (BPF). Like all filters, BPFs may be characterized by their amplitude and phase response versus frequency. A more tangible form of the latter characteristic can be derived by quantifying the change of phase divided by the change in frequency versus frequency. We call this characteristic group delay.

Due to the cost and complexity of its engineering, group delay is often ignored in favor of the amplitude response. This approach will soon be totally unacceptable as transmission rates in modern day communication systems are pushed to ever higher levels. While cost may still be an issue, the complexity of engineering group delay in a design is simplified by the fabrication techniques of SAW manufacturers. Because convenience often leads to ignorance, the design of a discrete UHF delay equalizer is presented as a tutorial for the interested engineer.

UHF Bandpass Filter

Figure 1 is the magnitude and delay

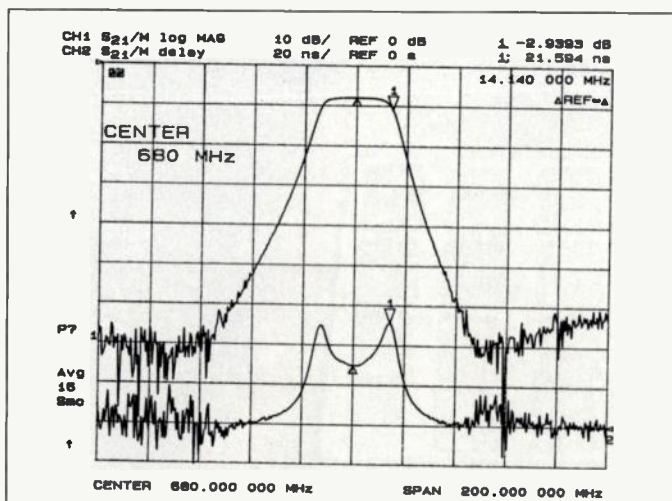


Figure 1. UHF BPF, magnitude and delay.

plot of power transmission through a UHF BPF. The filter was constructed with 11 air core inductors and 4 variable capacitors. With a passband ripple of 0.25 dB, a group delay ripple of 21 nsec, the filter has an insertion loss of 8.1 dB and a -3 dB bandwidth (BW) of 28.28 MHz. This BW is compatible with satellite television transmissions where modulating frequencies of up to about 4 MHz deviate carriers about 10 MHz. Frequency modulation (FM) of a carrier is well known to improve received signal-to-noise ratio (SNR) by spreading a given amount of power over a wider BW.

Because the FM process results in power being transmitted in sidebands that extend beyond the peak deviation of the carrier, the "Carson's rule" BW is often used in the reception of FM as a compromise between signal distortion and input noise reduction. The Carson's rule BW is given by:

$$BW(CR) = 2f_m(1 + M_p) \quad (1)$$

where M_p is the peak modulation index of the highest modulation frequency (or the peak deviation divided by f_m) and f_m is the highest modulation frequency.

Optimum demodulation of FM is largely influenced by the preselection filter i.e. the BPF. A frequency or angle demodulator is ideally insensitive to minor fluctuations in amplitude of the carrier. But at low signal levels amplitude changes affect all real demodulators. Therefore the passband should be flat. Adjacent FM transmissions are seen as noise by demodulators and therefore the BPF should have steep skirts to minimize all out-of-band noise. Steep skirts in the amplitude response of polynomial filters give rise to peaked responses in the delay of the same. Therein lies the dilemma, non-constant delay across a band of frequencies distorts their time relationship.

Delay Distortion

Most are familiar with the time delay and overshoot associated with a squarewave passing through a steep cut-off lowpass filter (LPF); fewer are familiar with the distortion caused by non-constant delay in wideband frequency demodulation. Figure 2 illustrates how delay distortion in the frequency domain is translated into amplitude distortion in the time domain.

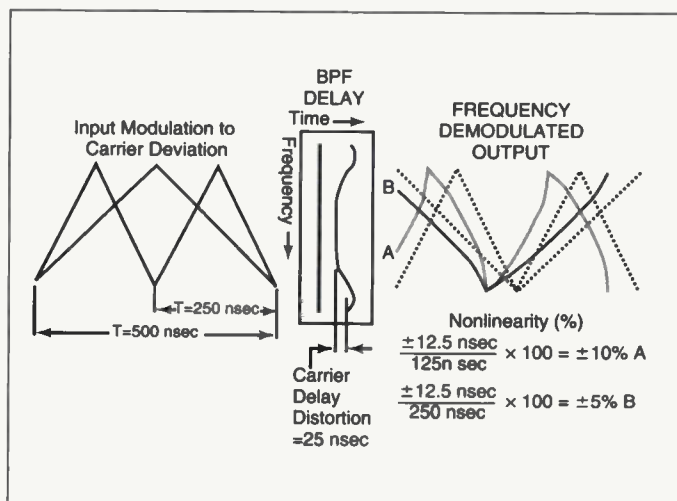
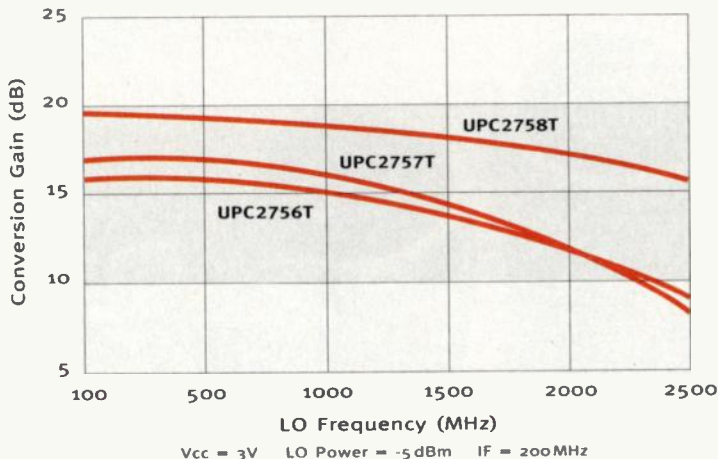


Figure 2. Delay distortion in an FM system.

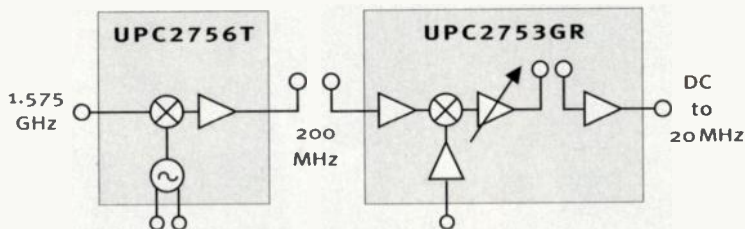
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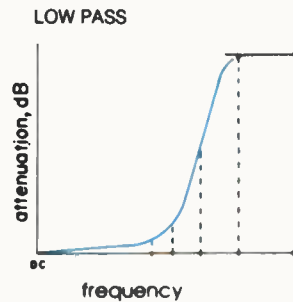
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Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	loss > 40dB
*LP-1.9	DC-1.9	3.4-4.7	4.7-200
*LP-2.5	DC-2.5	3.8-5.0	5.0-200
*LP-5	DC-5	8-10	10-200
*LP-10.7	DC-11	19-24	24-200
*LP-21.4	DC-22	32-41	41-200
*LP-30	DC-32	47-61	61-200
*LP-50	DC-48	70-90	90-200
*LP-70	DC-60	90-117	117-300
*LP-90	DC-81	121-157	157-400
*LP-100	DC-98	146-189	189-400
*LP-150	DC-140	210-300	300-600

dc to 1200MHz

Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	loss > 40dB
*LP-200	DC-190	290-390	390-800
*LP-250	DC-225	320-400	400-1200
*LP-300	DC-270	410-550	550-1200
*LP-450	DC-400	580-750	750-1800
*LP-550	DC-520	750-920	920-2000
*LP-600	DC-680	840-1120	1120-2000
*LP-750	DC-700	1000-1300	1300-2000
*LP-800	DC-720	1080-1400	1400-2000
*LP-850	DC-780	1100-1400	1400-2000
*LP-1000	DC-900	1340-1750	1750-2000
*LP-1200	DC-1000	1620-2100	2100-2500

All models priced qty. 1-9 (\$ea.), Conn. Type P = 11.45, B = 32.95, S = 34.95, N = 35.95
• Exceptions: *LP-1.9 P = 13.95, B = 34.95, *LP-2.5 P = 14.95, B = 35.95
On both models, add following to B price: \$3.00 for N, \$2.00 for S
75 ohm versions available

Surface-mount

dc to 108MHz

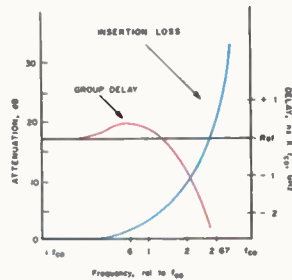
SCLF-5	DC-5.0	8-10	10-200
SCLF-8	DC-8.0	12.5-16.5	16.5-200
SCLF-10.7	DC-11	19-24	24-200
SCLF-21.4	DC-22	32-41	41-200
SCLF-25	DC-25	36-47	47-200
SCLF-30	DC-30	47-61	61-200
SCLF-45	DC-45	70-90	90-200
SCLF-95	DC-95	146-189	189-400

dc to 1200MHz

SCLF-135	DC-135	210-300	300-600
SCLF-190	DC-190	290-390	390-800
SCLF-225	DC-225	340-440	440-1200
SCLF-380	DC-380	580-750	750-1800
SCLF-420	DC-420	750-920	920-2000
SCLF-550	DC-550	800-1050	1050-2000
SCLF-700	DC-700	1000-1300	1300-2000
SCLF-1000	DC-1000	1620-2100	2100-2500

Price: SCLF 21.4-SCLF 420 \$11.45 ea. SCLF-8, 10.7, 550, 700, 1000 \$12.95 ea. SCLF-5 \$14.95 Qty. (1-9)

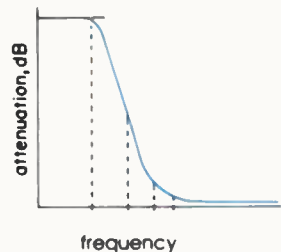
Flat Time Delay, dc to 1870MHz



Model No.	Passband MHz loss < 1.2dB	Stopband MHz loss > 10dB	loss > 20dB	VSWR Freq. Range, DC thru 0.2fco X	0.6fco X	Group Delay Variations, ns Freq. Range, DC thru fco X	2fco X	2.67fco X
*BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.70	4.0	5.00
*BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.90
*BLP-156	DC-94	312-416	416	1.3:1	1.1:1	0.30	1.1	1.50
*BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.40	1.3	1.60
*BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.20	0.6	0.80
*BLP-467	DC-280	934-1246	1246	1.25:1	2.2:1	0.15	0.4	0.55
*BLP-933	DC-560	1866-2490	2490	1.3:1	2.2:1	0.09	0.2	0.28
*BLP-1870	DC-850	3740-5000	5000	1.45:1	2.9:1	0.05	0.1	0.15

Price, (1-9 qty), all models: plug-in \$19.95, BNC \$36.95, SMA \$38.95, Type N \$39.95
NOTE: *BLP-933 and -1870 only with N and SMA connectors.

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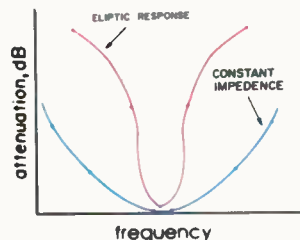
Model No.	Stopband MHz loss > 40dB	loss > 20dB	Passband, MHz loss < 1dB	VSWR Pass-band Typ.
*HP-25	DC-13	13-19	27.5-200	1.7:1
*HP-50	DC-20	20-26	41-200	1.5:1
*HP-100	DC-40	40-55	90-400	1.5:1
*HP-150	DC-70	70-95	133-600	1.8:1
*HP-175	DC-70	70-105	160-800	1.5:1
*HP-200	DC-90	90-116	185-800	1.6:1
*HP-250	DC-100	100-150	225-1200	1.3:1
*HP-300	DC-145	145-190	1290-1200	1.7:1

210 to 2200MHz

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

Price, (1-9 qty), all models: plug-in \$14.95, BNC \$36.95, SMA \$38.95, Type N \$39.95. For *HP-25, Add \$2 ea. floss 1.5 dB max.

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Model No.	Center Freq. (MHz)	Passband I.L. 1.5 dB Max. (MHz)	3 dB Bandwidth Typ. (MHz)	I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
*BP-10.7	10.7	9.5-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
*BP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
*BP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
*BP-60	60.0	55.0-67.0	49.8-70.5	44 & 79	4.6 & 190-1000
*BP-70	70.0	63.0-77.0	58.0-82.0	51 & 94	6.0 & 193-1000

Constant Impedance, 21.4 to 70MHz

Model No.	Center Freq. MHz	Passband MHz loss < 1dB	Stopband loss > 20dB at MHz	VSWR Total Band MHz
*IF-21.4	21.4	18-25	1.3 & 150	DC-220
*IF-30	30.0	25-35	1.9 & 210	DC-330
*IF-40	42.0	35-49	2.6 & 300	DC-400
*IF-50	50.0	41-58	3.1 & 350	DC-440
*IF-60	60.0	50-70	3.8 & 400	DC-500
*IF-70	70.0	58-82	4.4 & 490	DC-550

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

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A linear triangle wave linearly deviates a carrier. However, the spectral components of the modulated carrier are passed by a Carson's rule BW filter with different delays (energy at the band edges is delayed more). Though the demodulation process is linear-to-frequency, the demodulated signal is distorted and delayed relative to the unprocessed signal (dotted lines). Furthermore the distortion increases for increasing frequency input signals. This is readily seen if we define non-linearity in terms of the peak to peak time deviation over the sweep time (see Figure 2).

Inspection of the demodulated waveform leads to the conclusion that a greater portion of power is being shifted into the harmonics of the original waveform. This phenomena is intentionally used in some radar systems to produce a pulse output for a linear sweep input. These high slew transitions are where a PLL FM demodulator would be most likely to lose lock.

Though post-detection filtering could eliminate some harmonic distortion, the SNR of the fundamental signal is already reduced. Inspection also shows that by reducing the carrier deviation the incurred delay ripple could be reduced and thus the distortion could also be reduced. Unfortunately this results in less than optimum use of BW with a resulting reduction in SNR. Clearly some form of delay compensation is desired.

UHF Delay Equalizer

A second order delay equalizer (DE) ideally passes all frequencies unattenuated and produces a peaked delay at the design frequency. The peaked delay falls off at a rate proportional to

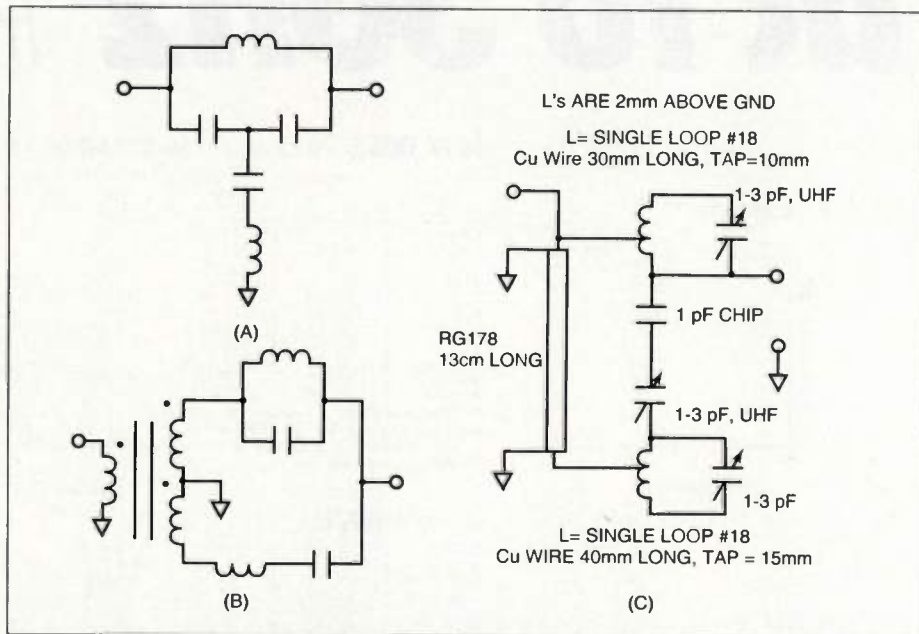


Figure 3. Second order delay equalizer a) Bridged T, b) Semilattice, and c) realization of equalizer.

the "Q" of the resonant network. The conventional approach to the realization of a second order DE is shown in Figure 3a. This bridged-T form works well at lower frequencies but results in low Q or impractical values at UHF.

Any chance of success at UHF requires the utilization of the more basic semilattice form shown in Figure 3b. This form is composed of a differential transformer, a parallel resonant and series resonant tank. The differential transformer performs the task of equally splitting or combining power through the resonant arms. It additionally provides 180 degrees of phase shift between the arms such that the transmitted power recombines constructively at the output. The

delay peak is dominated by the higher of the two resonant tank Q's, but when the Q's are equal is given by:

$$T(gd, \max) = (4Q)/\Omega_r \quad (2)$$

where Ω_r is the resonant radian design frequency.

A practical implementation of the semilattice UHF DE is shown in Figure 3c. A half wavelength of coaxial cable replaces the differential transformer and tapped resonant circuits are used to maximize component Q values. Tap points must be optimized empirically but the variable C's in the lower series resonant arm allow some post construction Q adjustment. Mutual coupling between the coils must be

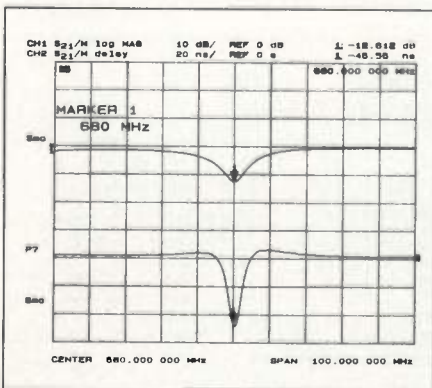


Figure 4a. Network response parallel arm.

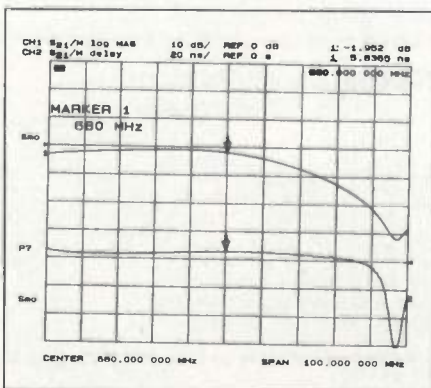


Figure 4b. Network response series arm.

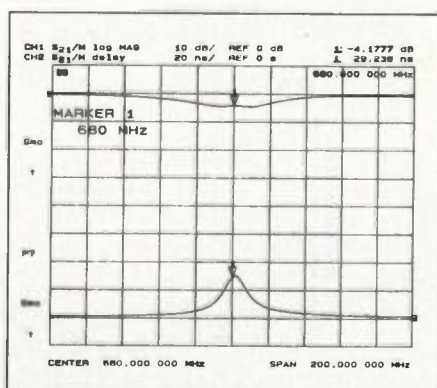


Figure 4c. Network response delay equalizer.

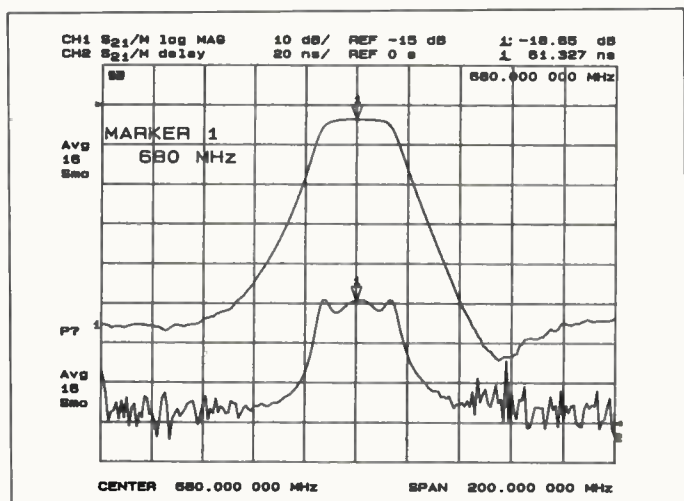


Figure 5. Cascade response — BPF, 5 dB pad, and delay equalizer.

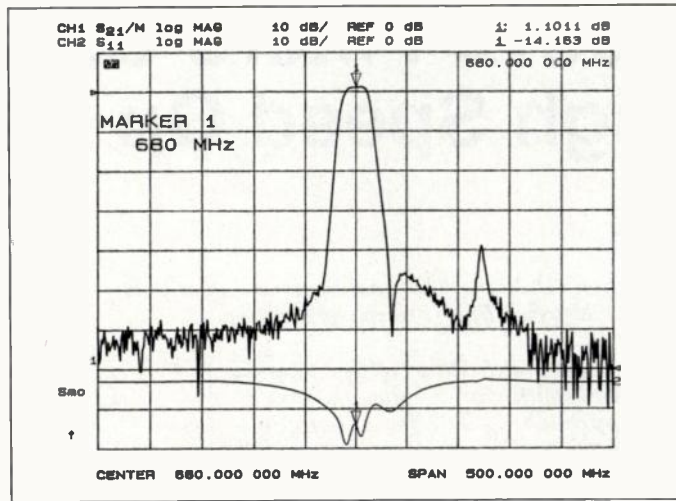


Figure 6. Amplitude response and delay of buffer, delay equalizer and BPF cascade.

avoided for proper adjustment.

Figure 4a is a magnitude and delay plot of the isolated parallel resonant arm and Figure 4b is of the isolated series arm. It can be seen that equal Q's were not realized, resulting in a passband ripple of 0.45 dB in the composite response of Figure 4c. The parallel arm, having higher Q, virtually determines the delay while the lower Q series arm is responsible (through feedback) for most of the insertion loss. The coaxial line was made shorter than half a wavelength, shifting its output-shorting-response to a parasitic resonance of the subject BPF.

Though optimized in a 50 Ohm system this DE requires a pad or buffer for cascading. A buffer amplifier could also compensate for the 4.2 dB insertion loss incurred. Alignment of the DE network on an analyzer is fascinating to watch as the delay flips from negative to positive.

Figure 5 is the composite plot of magnitude and delay of the BPF of Figure 1, a 5 dB 50 Ohm attenuator, and the DE of Figure 4c. The BPF was readjusted for best response. Passband ripple is still less than 0.5 dB, the -3 dB BW has been widened only slightly to 28.38 MHz, and the group delay ripple has been reduced to 6.7 nspp or ± 3.35 nsec. Commercially available SAW filters for satellite applications are often specified with a group delay ripple of ± 5 nsec. It should be noted that though the delay ripple has been reduced, the total delay has increased — this is an important consideration in demodulation circuits that involve frequency compression feedback as a

method of extending threshold.

A difference possibly worth noting is that this composite delay has three cycles of ripple across its passband where a SAW filter may have three times as many. This would create three times as many high slew transitions for a sweep across the filter with three times the probability of a PLL FM demodulator losing lock. Also the loss of the DE should not necessarily be considered a drawback as a clever person could use the concave amplitude response of a DE to flatten the convex response of a BPF!

Finally, Figures 6 and 7 show the amplitude response and delay for a cascade of a buffer amplifier, the delay equalizer, and the bandpass filter with a lowside notch added. The delay ripple is less than 3 nspp at a -3 dB BW of 26.7 MHz.

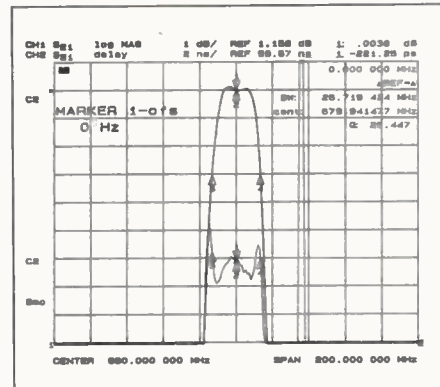


Figure 7. Close-in view of amplitude response and delay of buffer, delay equalizer and BPF cascade.

Conclusion

It has been illustrated how delay distortion of an RF carrier is translated into amplitude distortion through frequency demodulation. It has also been demonstrated that a polynomial filter with steep cut-off skirts has peaks in its delay response. If both low distortion and high SNR is desired in an RF system, delay compensation should be employed. Although at UHF frequencies an ideal DE was not achieved, delay compensation with negligible passband distortion was. **RF**

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1. Anatol I. Zverev, *Handbook of Filter Synthesis*, John Wiley and Sons, Inc., N.Y. 1967. chp. 3 & 10.
2. Jacob Klapper, John T. Frankle, *Phase-Locked and Frequency Feedback Systems*, Academic Press Inc., N.Y., N.Y. 1972, pp. 31-38.
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About the Author

Eugene Mayle is a senior engineer/project manager at the R.L. Drake Co. where he has been employed for the last 11 years. He received his BSEE from General Motors Institute in 1981 and his MSEE from The University of Dayton in 1986. He can be reached at (513) 746-4556.

Active Filters Using High Speed Op Amps

By Anthony D. Wang, National Semiconductor and
Kenneth Murray, Burr-Brown Corporation

The rapid proliferation of wideband amplifiers (with gain-bandwidth products approaching 1 GHz) has made active filter topologies more inviting as basic system building blocks. However, there are some configurations that are more readily suited for current-feedback amplifiers while others work best with voltage-feedback amplifiers. This article covers topological considerations, some practical construction tips and presents computer simulation techniques that simplify the design process.

Active filters using op amps have traditionally been limited to audio and industrial process control applications where the signals to be processed are (relatively) slow. Additionally, they helped to eliminate the need for inductors which can be quite large and cumbersome (not to mention expensive) at those low frequencies.

The advent of digital signal processing (DSP) and inexpensive analog-to-digital converters has reduced the implementation of active filters at lower frequencies (<100 kHz). However, new op-amps have arrived which allow the implementation of active filters at frequencies which DSP filtering is prohibitively expensive.

At the same time, competition in the op amp market has driven prices down — for example, the OPA4658, a quad 900 MHz op amp, has a 100 piece price of \$7.55, or less than \$2 per amplifier! These performance and economic considerations make op amp active filters suitable for high frequency applications.

Amplifier Architectures

One aspect of active filter implementation that requires consideration is the selection of the op amp type: Current-Feedback Amplifier (CFA) or Voltage-Feedback Amplifier (VFA). Traditional filter topologies were designed with the VFA as the active element. Although the CFA uses the same element symbol as the VFA, there are

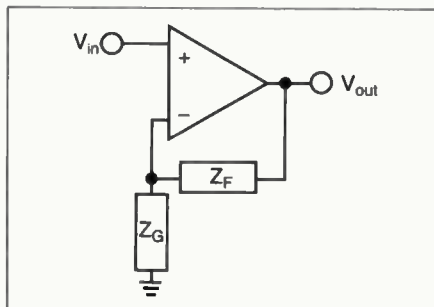


Figure 1. Op amp with feedback.

some differences that prevent its being used as a generic replacement.

Why then bother with the current-feedback amplifier if the voltage-feedback variety with similar bandwidths is available? This question has merit. However, a brief review of the salient differences will hopefully provide the insight needed to make the proper selection.

Briefly, the CFA has the property of providing a closed-loop bandwidth that is (relatively) independent of its closed-loop gain. The VFA is limited by the gain-bandwidth product (GBW) which is the traditional figure of merit used for comparison.

Another characteristic that differentiates the two amplifier types is the large signal driving capability of the CFA. Current-feedback amplifiers have very good slew rate properties when compared to voltage-feedback amplifiers and this trait is very beneficial if large signal distortion is to be minimized.

The CFA does not violate any physical laws to achieve its properties; the key to understanding it is to realize that it is an uncommitted amplifier block whose open-loop gain is set by the external impedance connected to the inverting input node.

The VFA, on the other hand, has its open-loop gain set by design and the external elements do not affect it. [1]

For the CFA, the same external elements that set the closed-loop gain

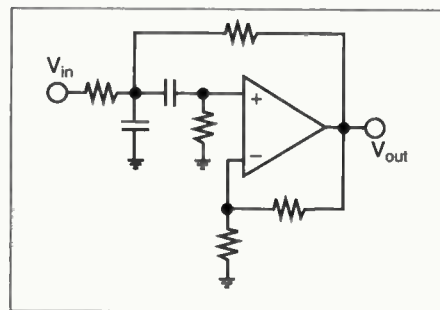


Figure 2. Sallen-Key filter.

(Figure 1) contribute to the open-loop gain according to equation 1.

$$A_V = \frac{R_T}{Z_{EQ} + R_{IN}} \cdot \frac{1}{1 + j\omega R_T C_T} \cdot e^{-j\omega T_D} \quad (1)$$

where

$$Z_{EQ} + Z_F || Z_G$$

The parameters R_T , C_T , and R_{IN} are inherent to the amplifier and differ from one product type to another (much as GBW for VFAs differ). T_D is a mathematical artifact that is used to approximate the excess phase in the open-loop response.

If the feedback network is purely resistive, the open-loop response is easy to analyze.

$$A_V = \frac{R_T}{\frac{R_F \cdot R_G}{R_F + R_G} + R_{IN}} \cdot \frac{1}{1 + j\omega R_T C_T} \cdot e^{-j\omega T_D} \quad (2)$$

If either, or both, of the elements in the feedback network is reactive with the addition of parallel capacitors, for instance — the open-loop behavior becomes more complex (pun is purely intentional!).

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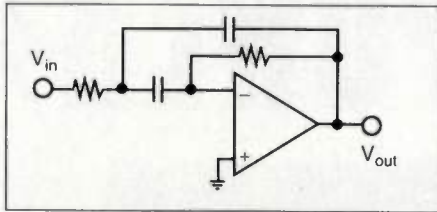


Figure 3. Negative feedback bandpass filter.

$$A_v = \frac{R_T}{R_{EQ} + R_{IN}} \cdot \frac{(1 + j\omega R_{EQ} C_P) \cdot e^{-j\omega T_D}}{(1 + j\omega R_T C_T) \left(1 + j\omega \frac{R_{EQ} \cdot R_{IN}}{R_{EQ} + R_{IN}} C_P \right)} \quad (3)$$

Equation 3 shows the resulting open-loop gain expression for this situation. The parallel combination of the feedback network resistors has been relabeled R_{EQ} and the net parasitic capacitance has been lumped into C_P .

The altered response has added a zero and a pole, but the zero will always occur before the pole. This can cause stability problems and is the reason that CFAs are typically troublesome in integrating applications.

The basic points to keep in mind are that the CFA is optimally used for high gain and/or fast, large-signal situations. The VFA is best for integrating applications. Small-signal situations with low closed-loop gains could be satisfied by either.

Active Filter Topologies

The available topologies for active filters are too numerous to cover in this article but some well documented configurations are presented here to illustrate the ideas presented above, namely how to select the appropriate op amp type. The discussion will focus on the biquadratic filter function which is best described as the ratio of two second order functions.

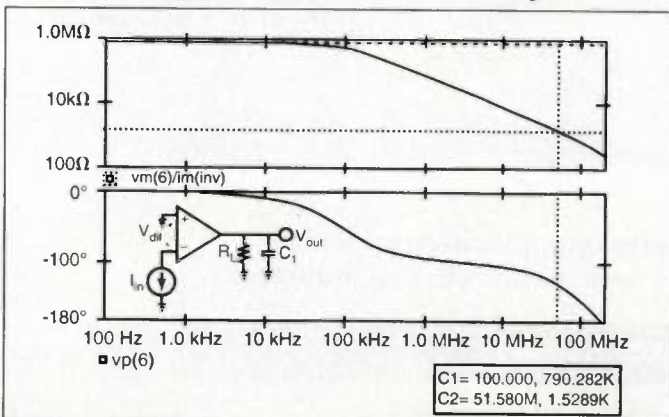


Figure 5. Measuring open-loop transimpedance (top), and open-loop phase (bottom).

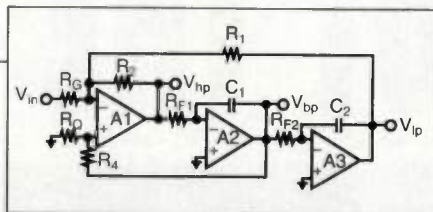


Figure 4. State-variable active filter.

This is often used as the basic building block for higher order filters because it allows precise placement of the complex conjugate pairs. By cascading these second order sections, higher order filters can be easily realized without the associated problem of interacting pole pairs. Each section can be designed and tuned independently.

Figure 2 illustrates the popular Sallen-Key band-pass configuration. Note that this circuit uses positive feedback. The Q of this circuit is set by

$$Q = \frac{1}{3 - K} \quad (4)$$

where K is the closed-loop gain of the op amp.

If the amplifier is configured for unity gain, the Q is 0.5 and a VFA would be the appropriate choice. (Most CFAs perform poorly in unity gain.) For a Q of 0.667 the amplifier has to be configured for a gain of 1.5 which, for a VFA, means a bandwidth reduction of 33 percent! Clearly the CFA can be much more useful in this application even though the closed-loop gain is not very high.

The drawback to this positive feedback filter circuit is that it is sensitive to component tolerances and gain accuracy and thus it is confined to applications with Q less than five.

Figure 3 shows an active filter configuration that utilizes negative feedback giving it a lower sensitivity to component tolerances. The net imped-

ance seen at the inverting input is complex which makes it a bit tricky to utilize the CFA in this application. The obvious choice here is to use a VFA.

Unfortunately, the negative feedback filter requires larger closed-loop gain which, for a VFA, means a reduction in usable bandwidth. This illustrates once more the penalty of the GBW limit for voltage-feedback amplifiers.

For high- Q applications, the multiple op amp circuit of Figure 4 is a more practical implementation because (1) it has lower sensitivity to the active and passive components, (2) Q and ω_0 can be adjusted independently, (3) the spread of component values is lower and (4) the gain requirements are less critical.

This circuit is also known as the state-variable (universal) active filter. The three separate op amps give the low-pass, band-pass and high-pass responses without having to be reconfigured. Wideband applications can be built using a combination of VFAs and CFAs.

As previously stated, the integrator is best implemented with the VFA while the CFA is generally the proper choice for the summing amplifier, A1.

Computer Simulation

Once the active filter topology has been determined, the design can be optimized using PSPICE (or some other program). The software is not as critical as are the models of the amplifiers being considered — the macromodels.

Macromodels are circuit descriptions of the op amps that are supplied (usually) by the vendor. They have been simplified to make simulations faster, usually by minimizing the number of active components (i.e., the transistors) in the listing. In high speed designs, the

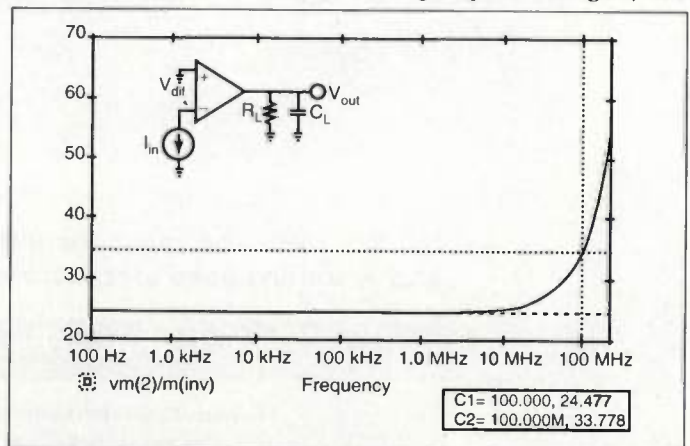


Figure 6. Measuring input impedance.

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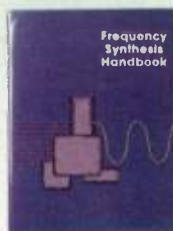
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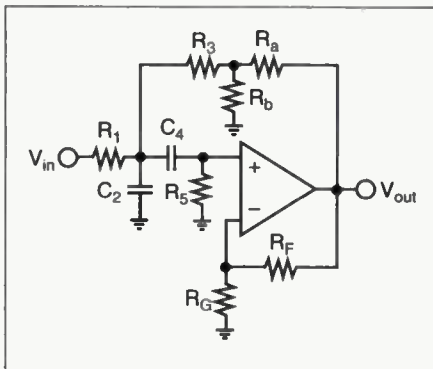


Figure 7. Modified Sallen-Key bandpass filter.

degree of simplification is important.

For low frequency designs, the load resistances are generally higher because the associated RC time constants do not typically limit the bandwidth of the system. Therefore the amplifier output stages do not have to supply large variations in current. The macromodels for these op amps usually consist of two transistors to accurately model the input stage while the rest of the amplifier is composed of passive components and controlled sources. From the standpoint of simulation, this type of circuit is computationally efficient.

This is not true for wideband designs which have another penalty because parasitic capacitances cannot be neglected. The amplifiers have to drive more current at higher speeds and still minimize distortion. Therefore, to accurately model these op amps, more transistors are needed in the macromodel. These are generally limited to just the high speed signal path and output stage. The tradeoff is a longer simulation time for a higher confidence level in the results.

Once the appropriate macromodels are found, simulations can be run. The passive component values can be determined with the many filter programs that are presently available (e.g., FilterPro which is free from Burr-Brown). Actual circuit implementation is straightforward as long as only VFAs are used. If a CFA is to be used (for reasons stated earlier), then there is a little more work involved.

It is usually assumed that by proper selection of the feedback resistor, the CFA can be tamed and configured properly. This is true, but without a clear methodology, the user can only home in on an optimum value by trial and error. Fortunately, after the proper value is found, straightforward impedance scaling can be used to change the other passive component values so that filter performance is unaffected.

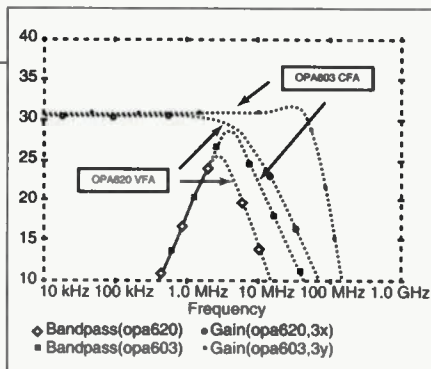


Figure 8. Sallen-Key simulation results.

There is a direct approach that makes the resistor selection painless. As stated previously and shown in equation 1, the current-feedback amplifier has its open-loop response determined by intrinsic characteristics and the impedance at the inverting input node. Although these intrinsic parameters are not always available from the product datasheet they can be determined from simulation.

Figure 5 shows the basic approach and the corresponding response curves for a generic CFA. The term "current-feedback" means that the amplifier is responsive to a current at its inverting input. Therefore, the proper stimulus is a current source. The open-loop response has the shape that is normally associated with VFAs but the units are in ohms (volts out/amps in) and it is referred to as transimpedance.

R_T , the transresistance, is the dc value of transimpedance. C_T , the transcapacitance, interacts with R_T to cause the open-loop rolloff. It is determined by noting where the open-loop phase has fallen 45°. The value of T_D is calculated to closely approximate the excess phase at some critical point, 60° phase margin, for instance.

R_{IN} is the dc value of the impedance looking into the inverting input. Figure 6 illustrates the simulation technique for its measurement. Although the inverting input impedance is complex, the dc portion is dominant over the CFA's useful bandwidth.

The PSPICE listing for the simulation is detailed below:

```
* CURRENT-FEEDBACK OPEN-LOOP SIMULATION *
* file: CFA-OL.CIR
*****Simulation Commands *****
.options noecho nomod numdgt=8
.op
.ac dec 20 100 500meg
.probe
***** Library Files *****
.lib BURR_BRN.LIB
*****Circuit Listing *****
vp 7 0 5
vm 4 0 -5
vin 16 0 dc 0 ac 1
xf1 16 3x 2x opa620 filter
xoal 3x 2x 7 4 opa620 opa620x/bb
xf2 16 3y 2y opa603 filter
xoal 3y 2y 7 4 opa603 opa603/bb-x
***** filter model *****
```

```
*** inv is the excitation source for open-loop analysis ***
inv 2 0 dc 4.097ua ac 1
xcfa 0 2 7 4 6 opa603/bb-x
rl 6 0 100
cl 6 0 10pf
.end
```

Finding the right resistance to use in the feedback network can be accomplished mathematically, by using equation 2, or graphically, from the open-loop curves. The trick is to determine how much excess phase the design can tolerate. Once the phase criterion is known, the frequency, f_i , at which it occurs sets the open-loop transimpedance of the amplifier at that frequency.

The desired closed-loop gain should match the open-loop gain at the frequency which was determined by the phase criterion. Equation 5 gives the simplified expression.

$$\frac{R_F + R_G}{R_G} = \frac{|Z_T(f_i)|}{R_{IN} + R_{EQ}} \quad (5)$$

$$= \frac{R_T}{1 + j2\pi f_i R_T C_T}$$

$$= \frac{R_T}{R_{IN} + R_{EQ}}$$

Solving this expression gives the appropriate feedback and gain setting resistor values. There could be situations where the computed resistor values are very low or even negative. In these situations, a lower frequency design (higher transimpedance) has to be adopted.

Design Examples

Figure 7 is a modified version of the Sallen-Key bandpass filter shown in Figure 2. The circuit is configured for a pass-band gain greater than that required by the filter design equation. The R_a - R_b resistive divider is used to bring the amplifier output back down to the required level.

The passive component values in the PSPICE listing below were selected for 1 percent resistor tolerances and 5 percent capacitor tolerances.

```
* SALLEN-KEY BANDPASS FILTER *
* file: SK.CIR
***** Simulation Commands *****
.op
.ac dec 20 10k 200meg
.probe
***** Library Files *****
.lib BURR_BRN.LIB
***** Circuit Listing *****
vp 7 0 5
vm 4 0 -5
vin 16 0 dc 0 ac 1
xf1 16 3x 2x opa620 filter
xoal 3x 2x 7 4 opa620 opa620x/bb
xf2 16 3y 2y opa603 filter
xoal 3y 2y 7 4 opa603 opa603/bb-x
***** filter model *****
```



```
.subckt filter 16 3 2 6
r1 16 9 562
c2 9 0 62pf
r3 10 9 280
c4 9 3 62pf
r5 3 0 1.13k
ra 6 10 5.62k
rb 10 0 294
rf 6 2 681
rg 2 0 20
ri 6 0 150
.ends filter
.end
```

The simulation compares the design with two separate amplifiers. The OPA603 is a 60 MHz CFA while the OPA620 is a 200 MHz VFA. The resulting response is shown in Figure 8. The design with the OPA620 falls short because of its gain-bandwidth limitation while the design using the lower bandwidth OPA603 performs as expected because its closed-loop bandwidth has not suffered.

The state-variable filter of Figure 4 is simulated next. The listing below is for a second order Bessel highpass filter that was originally designed for -3 dB at 10 kHz. The "rmod" and "cmod" parameters provide easy frequency and impedance scaling capability to move the design up to 10 MHz.

```
* WIDEBAND UNIVERSAL ACTIVE FILTER SIMULATION *
* file: UAF.CIR
***** Simulation Commands *****
.options noecho nomod
.ac dec 40 1meg 1000meg
.probe
***** Library Files *****
.inc opa642x.mod
.inc opa644x.mod
.model rmod res(r=0.005)
.model cmod cap(c=0.2)
***** Circuit Listing *****
vp 11 0 5
vm 10 0 -5
vin 2 0 dc 0 ac 1
xflt1 2 opa642 3x 12x 11 10 uaf
xoa1 12x 3x 10 10 opa642 11 11 opa642x
xflt2 2 opa644 3y 12y 11 10 uaf
xoa2 12y 3y 10 10 opa644 11 11 opa644x
**** active filter subcircuit ****
.subckt uaf 2 13 3 12 11 10
*node list: 2=in 13=hp 7=bp 1=lp 3=+in 12=-in
rg 2 12 rmod 50k
rf1 13 8 rmod 20.5k
rf2 7 14 rmod 20.5k
r1 12 1 rmod 50k
r2 12 13 rmod 50k
r4 3 7 rmod 50k
rg 3 0 rmod 68.1k
c1 8 7 cmod 1000pf
c2 14 1 cmod 1000pf
xoa2 8 0 10 10 7 11 11 opa642x
xoa3 14 0 10 10 1 11 11 opa642x
.ends uaf
.end
```

The simulation uses the OPA642, a 450 MHz VFA, for the integrating amplifiers and compares the response with two different summing amplifiers: the OPA644, a 500 MHz CFA, versus the OPA642 VFA. As the simulation plot in Figure 9 shows, the CFA should deliver performance out to almost 300

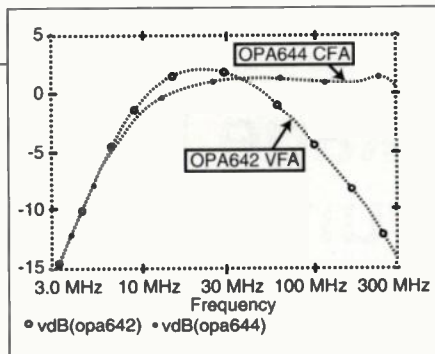


Figure 9. Simulation results for state-variable filter of Figure 4 using a CFA and VFA.

MHz with less than 0.5 dB ripple while the VFA does not work at all.

Figure 10 shows the measured results of actual circuits built using the schematic of Figure 4. The scan from 30 MHz to 500 MHz confirms the simulation prediction that the VFA's response would roll off well before that of the CFA. The bump that shows up on both traces is a board parasitic.

Construction Tips

High frequency board layout techniques are obviously needed for wideband active filters. There are some special considerations that need to be taken with high speed op amps.

Surface mount components should be used wherever possible. This is especially true for the high frequency decoupling capacitors which should be mounted as close to the supply pins as possible.

As equation 3 indicated, any capacitance at the inverting input of a current-feedback amplifier can alter its open-loop response. It also changes the characteristics of the feedback network and would thus be detrimental to voltage-feedback amplifiers as well. It is usually good practice to remove the ground plane around the op amp input pins. The connection of passive components to these pins should be made as close to the pins as physically possible.

Many wideband amplifiers provide separate power pins for the output stage. If these are available, run separate power supply traces to the pins rather than shorting the pins at the package. This has been shown to provide the lowest distortion performance.

Use the tips that the product datasheets offer. The vendor's design, test and applications engineers have usually travelled the hard road to make their products manufacturable. Many datasheets provide layouts of demo boards that were optimized for the associated op amp.

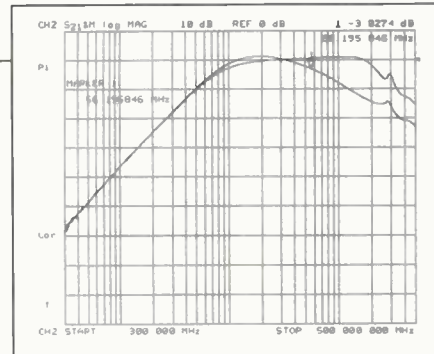


Figure 10. Measured results of state-variable filter of Figure 4 using a CFA and VFA.

Don't give up! Many high speed designs are iterative and involve finding every performance robbing parasitic. You can shortcut part of this process by trying to simulate every parasitic you can identify.

Conclusion

The availability of low cost, high speed op amps makes wideband active filters a viable alternative in the designer's bag of tricks. Proper selection of the op amp type — current-feedback vs voltage-feedback — will pull the highest performance from your design. **RF**

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About the Authors

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Simulator Package Models a Spread Spectrum System

Part 1: The Models

By Stephen Kratzet
ELANIX, Inc.

SystemView by ELANIX, Inc. is a dynamic system simulator software package that runs on 386/486 and higher PCs operating with Microsoft Windows. While SystemView is often used for simulating DSP systems that include digital multipliers, FIR filters, multi-rate sampling, and feedback paths that allow adaptive loop modeling, it is also an excellent tool for the Analog/RF designer. Here one is interested in the mixer spurs that a certain frequency plan may generate, how various filter types may delay the signal path, or how a system will respond to noise at its input.

Part 1 of this two part article will discuss three building blocks — a mixer, phase detector, and a biphasic modulator. The building blocks are not intended to have the detail of a SPICE subroutine but are similar to analog behavioral modeling [1]. Part 2, which will appear next month, will use the building blocks to simulate an entire transmitted reference spread spectrum system.

RF Mixer

This building block (Metasystem) approximates the operation of a mixer over a reasonable range of input amplitudes with its output containing mixing spurs, and RF/LO port leakage. The Metasystem, with its In and Out tokens does this in the following manner (Figure 1): Each of the inputs to the mixer is passed through a Function Limiter. With the limiters' Min Input and Max Output values equal to each other, a signal with an amplitude less than the limit value will pass through unchanged while amplitudes larger than the limits are clipped. The gain adjusted LO signal and the RF signal are fed into a 4-quadrant multiplier. The output of the mixer is the

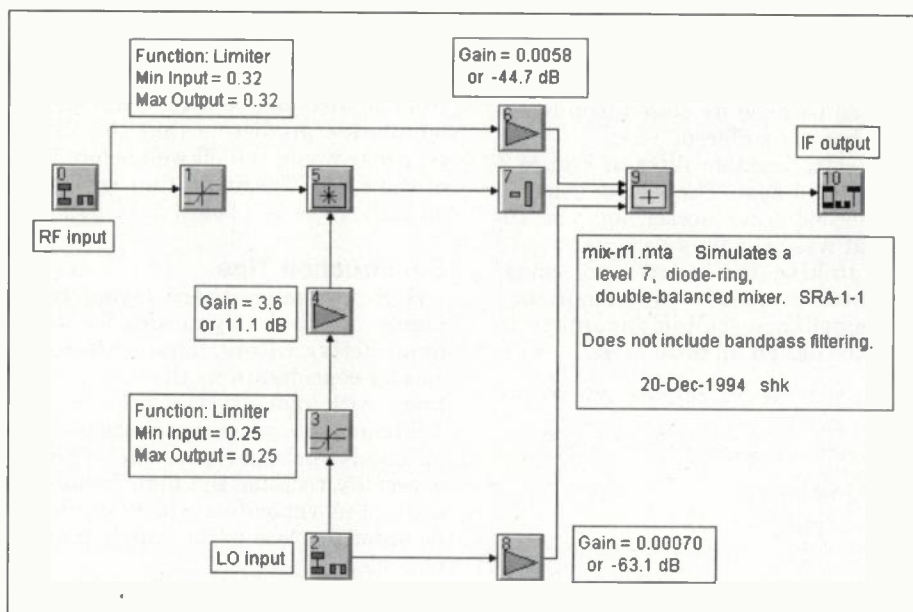
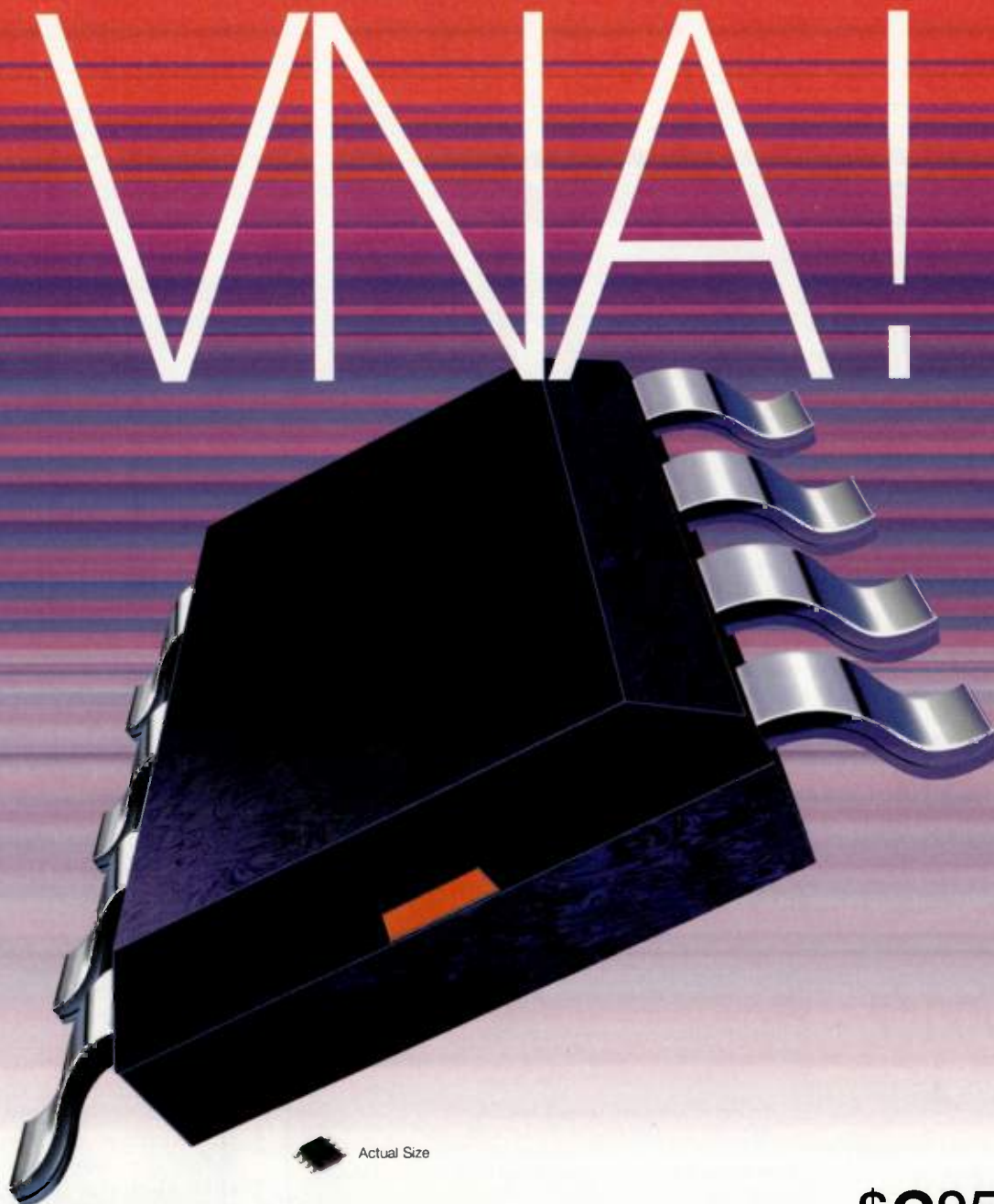


Figure 1. Mixer metasystem.

sum of the multiplier output and the 2 leakage paths. To be consistent with a hardware mixer that has a negative output when both inputs are the same phase, the multiplier output is inverted. This mixer model has no provision to limit the mixer's frequency response.

The value in the GAIN token may be entered as a linear gain or in dB. This allows custom mixers to be created and modified. A limitation of the model is the mixer performs mixing no matter how weak the LO becomes. Figure 2 shows the mixer Metasystem being fed by two sine waves and its output going to a "sink" for viewing. SystemView allows rapid building of simulations. Just double-click the mouse on the selected token family from the left edge of the window. Double-click again to define the type of token within the family, and fill in a

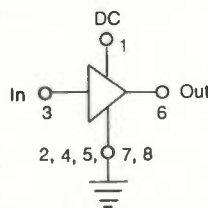
few parameters — amplitudes, frequencies, phases, and gains. The Metasystem may have been created and saved as a file earlier, or a Metasystem may be generated quickly by using a keyboard-Ctrl, mouse-click-and-drag to outline a group of tokens. SystemView runs as a time sampled system. A click of the mouse on the CLOCK button at the top center of the screen allows the setting of the System Time parameters. In this simulation 4096 samples at 204.8 Mega-samples per second (MSPs) is used to allow for the mixer spurs. The use of power-of-two numbers avoids the bin-splitting effects when the FFT is performed and gives a cleaner display for this particular test. To the left of the CLOCK button is the EXECUTE button, which is used to run the simulation. The button to the right of the CLOCK button is the "OSCILLOSCOPE button" that



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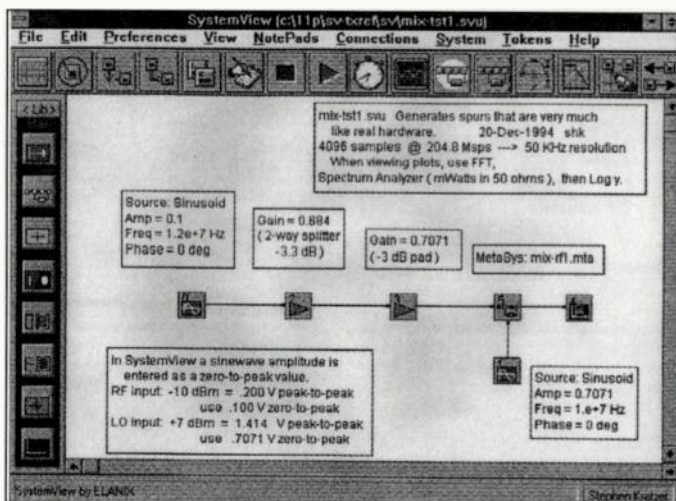


Figure 2. Mixer test system.

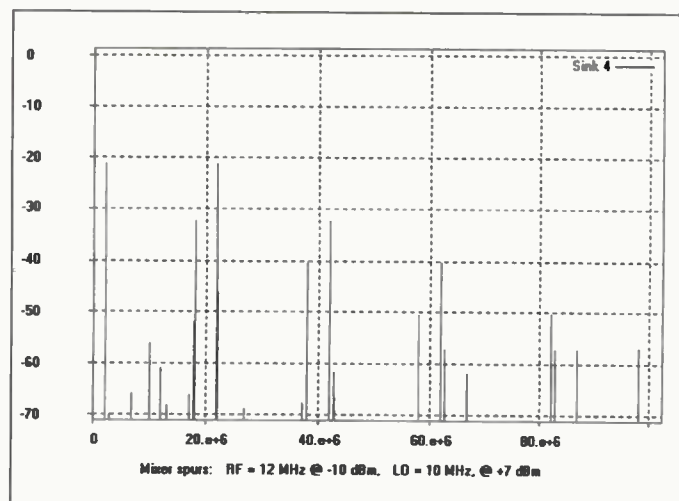


Figure 3. Mixer test system output plot.

opens the Analysis window for viewing the results of the simulation contained in the "sink" token. SystemView has an alternate way of displaying the buttons on either side of the CLOCK. By selecting Preferences, Toolbar, the "text buttons" will be replaced with

"graphic buttons" and vice versa.

Figure 3 shows the mixer spurs. The two main sum and difference frequencies of 22 MHz and 2 MHz are each at -22 dBm. This is due to the -10 dBm input being reduced by the -3.3 dBm splitter, -3 dBm pad, and a 5.7 dBm

loss through the mixer. Notice the RF port leakage, -62 dB at 12 MHz, and LO port leakage, -57 dB at 10 MHz. This plot is obtained in a few mouse-clicks: New Data, FFT, Spectrum Analyzer (mWatts in 50 ohms), then Log y. A keyboard-Ctrl, click-and-drag with the mouse allows a close inspection an area of interest on the display.

The parameters used in the mixer Metasystem are the result of measurements taken from a real-world hardware I-Q down conversion system. Figure 4 shows the SystemView model of an I-Q down converter, with a pair of Bessel lowpass filters. A filter may be created in SystemView with just a few steps: Double click the mouse on the Operator token family at the edge of the window, double-click again to define the token as a Linear-system, then click the following buttons — Parameters, IIR, Filter library, Bessel, 5 (the number of poles), Lowpass, enter 2e6 (2 MHz) for the Cutoff frequency. In a few moments the Time (impulse) response of the newly created filter will appear in a window. The Gain, Phase, Group Delay, Root Locus, or Bode plot (Figure 5) may also be viewed. A "Duplicate Tokens" feature allows the copying of a single token or a selected group of tokens.

The operation of the down converter in Figure 4 may be tested with various inputs. When the RF is -10 dBm and the LO is +7 dBm, and they are set to the same 10.0 MHz frequency and phase, the I output is zero volts, and the Q output is a constant -27 mV.

In Figure 6 the RF frequency has been changed to 10.05 MHz causing the I (In-phase) and Q (Quadrature)

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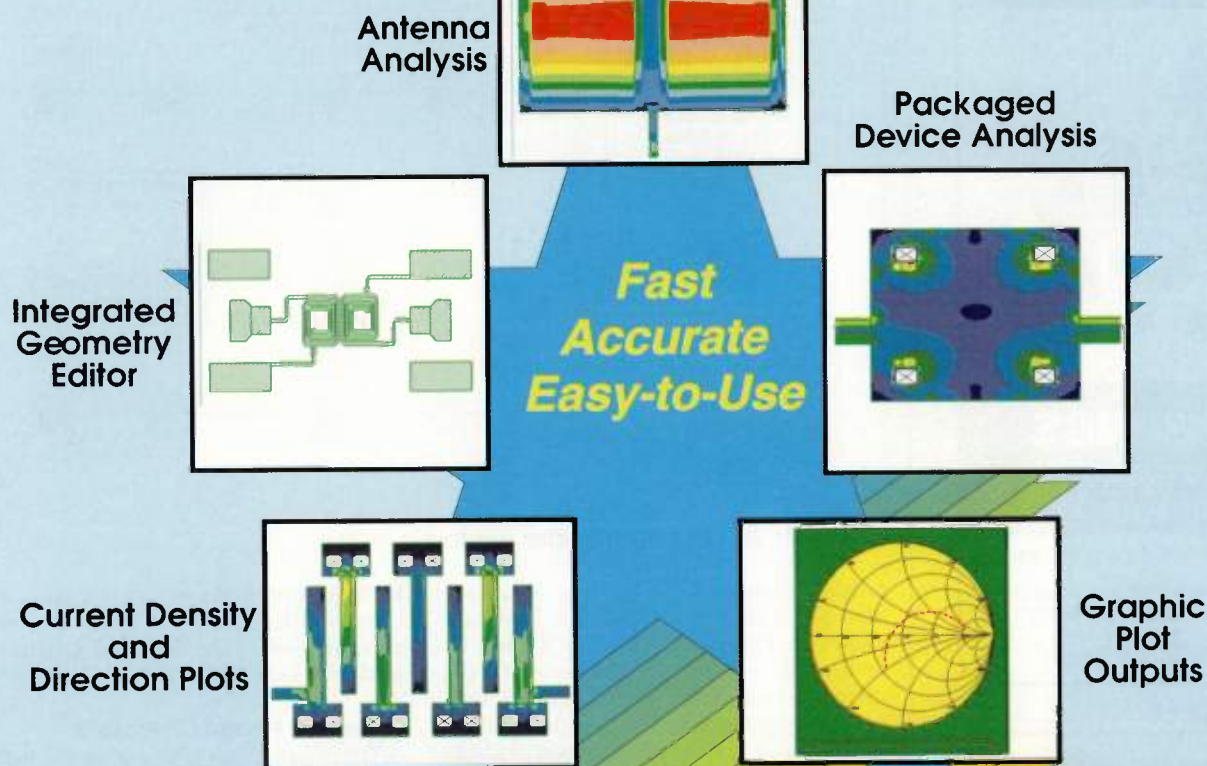
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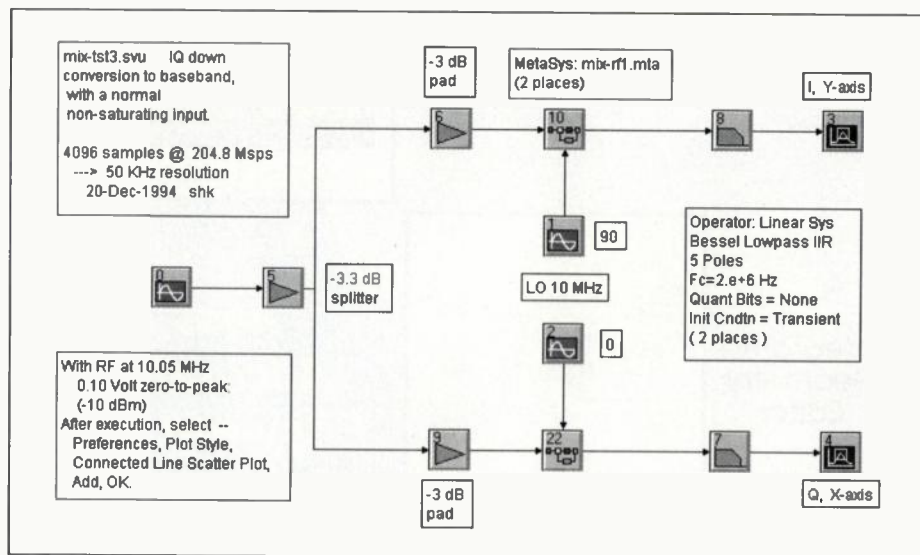


Figure 4. IQ Down converter system.

outputs to be 50 kHz. SystemView can also plot the X vs Y (Scatter Plot) as shown in Figure 7. With the RF input at a relatively low input power of -10 dBm the mixers are operating in a linear mode and produce a circular plot.

Figures 8 and 9 show a saturated

mode of operation when the RF input is raised to +13.3 dBm (1.462 volts, zero-to-peak) (the RF and LO mixer ports are both operated at +7 dBm).

Types Of Phase Detectors

There are three types of phase detec-

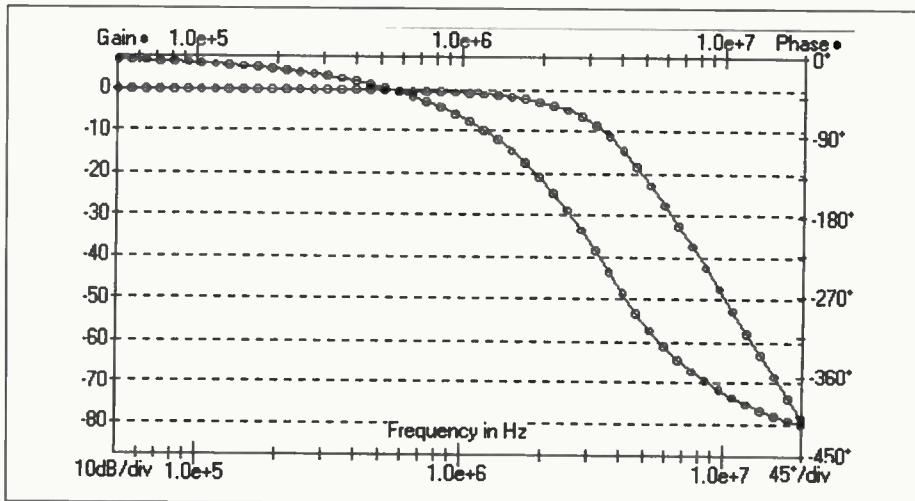


Figure 5. Bode Plot of 2 MHz lowpass filter.

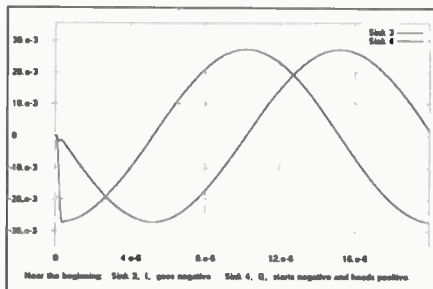


Figure 6. RF input 50 kHz higher than the LO input, time plot.

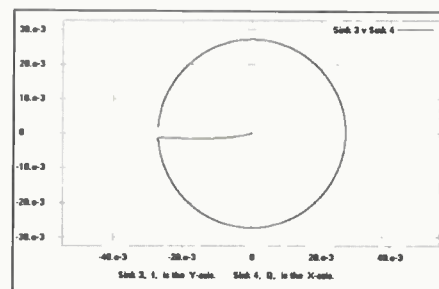
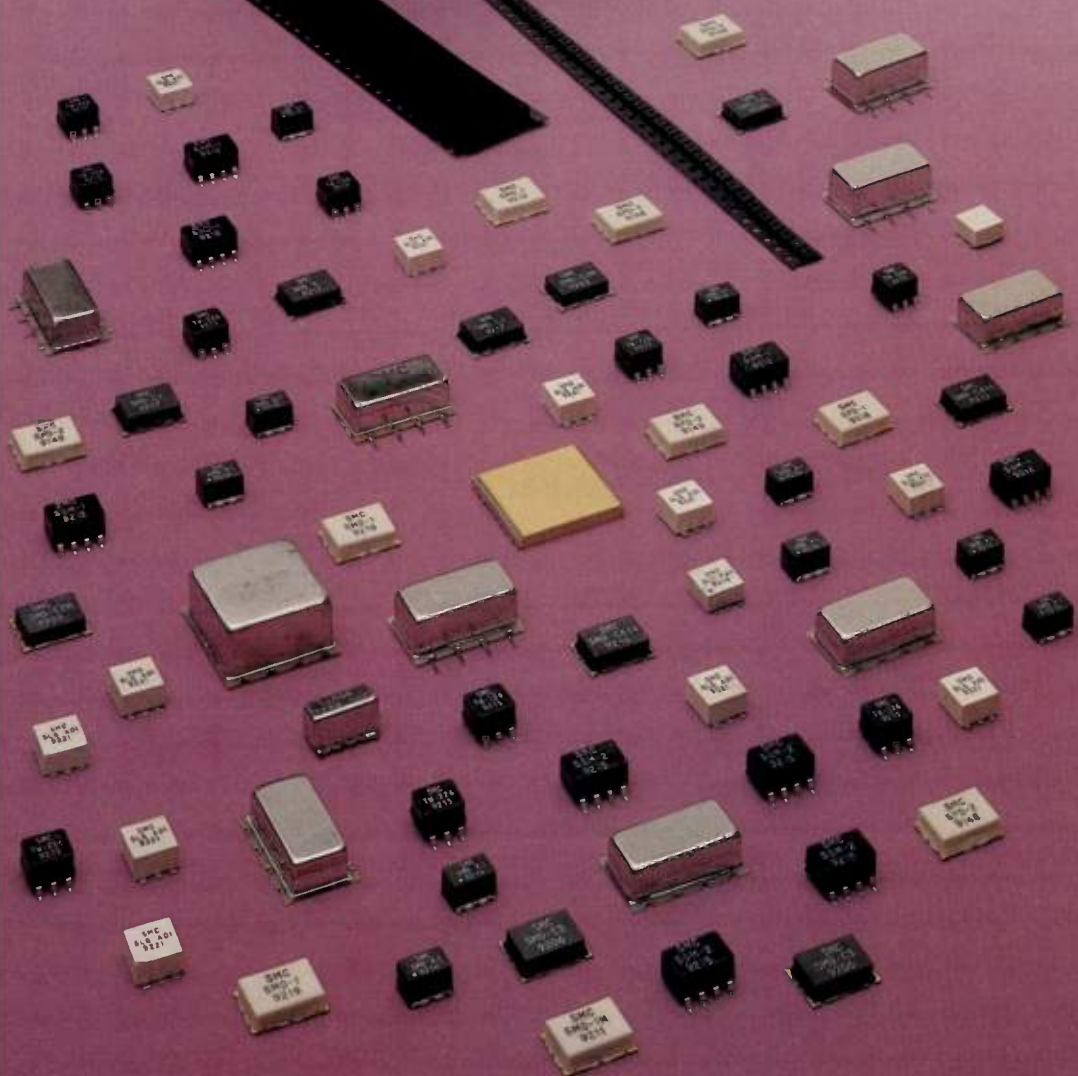


Figure 7. RF input 50 kHz higher than the LO input, scatter plot.

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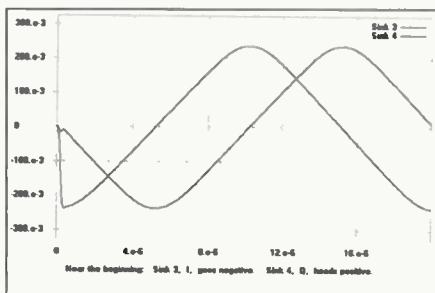


Figure 8. RF input 50 kHz higher than the LO input, time plot.

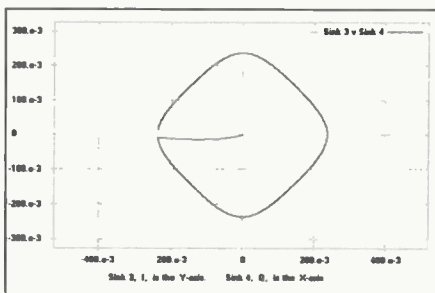


Figure 9. RF input 50 kHz higher than the LO input, scatter plot.

tors [2]: (1) Analog or multiplier (mixer); (2) Exclusive-OR (digital equivalent of the analog multiplier); and (3) Digital phase-frequency detector (CD4046 or MC4044 type). The main advantage of the analog phase detector is its ability to recover a signal from a low signal to-noise input. Unfortunately, it will also lock in on harmonics of the desired input. For the noise free TTL signals found in a frequency synthesizer, the third type of phase detector is a better choice. The CD4046 responds only to the rising edges of the inputs, eliminating the harmonics lock-in problem. (The

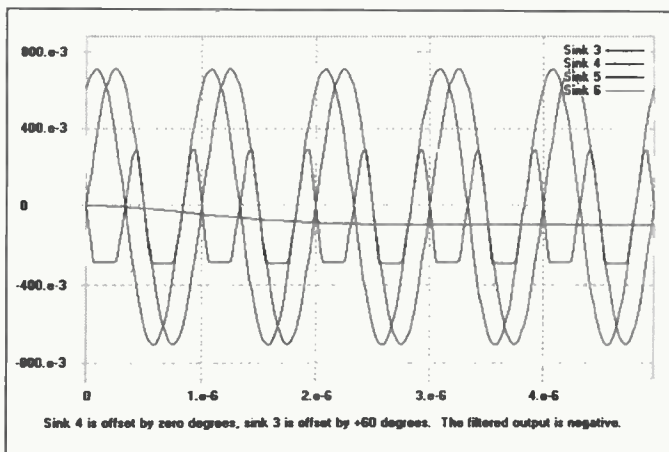


Figure 11. Two 1 MHz signals, 60 degrees apart, yields a negative output.

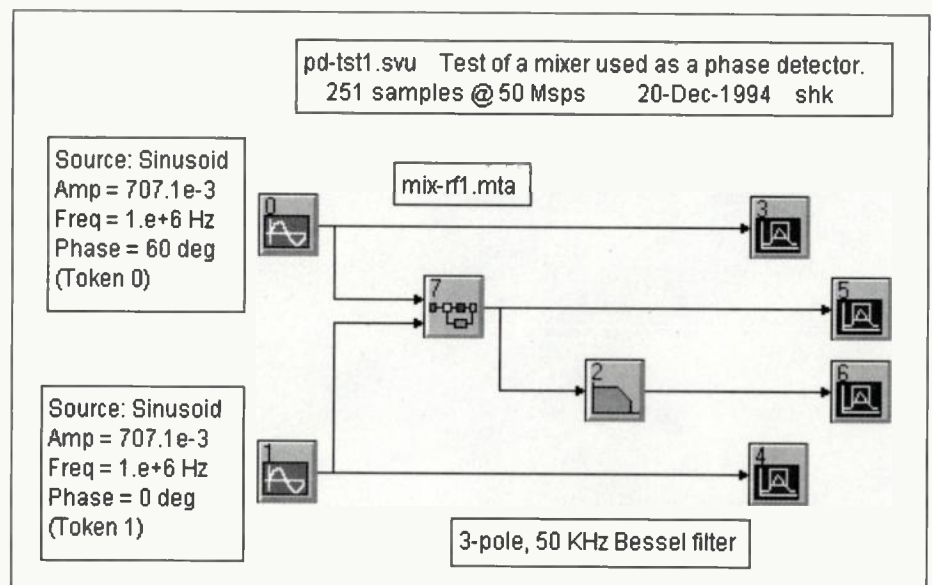


Figure 10. Mixer operated in the saturated mode as a phase detector.

MC4044 responds to the falling edges only.) The digital phase detector is not perfect, an extra or missing pulse generates a large error for a short time, and the polarity of its feedback connection is important.

Analog Phase Detector

By operating the mixer in a saturated mode it can perform analog phase detection without being amplitude sensitive [3] (Figure 10). When both inputs are in-phase (zero degrees) the output will be at its maximum negative value. With the inputs at ± 180 degrees phase difference the output will be at its maximum positive value. With the inputs at ± 90 degrees phase difference the output will be zero. Figure 11 shows the phase detector out-

put for a phase error of 45 degrees. When this type of phase detector is used in a phase-locked-loop (PLL), the loop will lock-in regardless of the feedback polarity.

Biphase Modulator

For testing and comparison a hardware version of a modulator was built as shown in Figure 12 [4].

The Metasystem created for this modulator approximates the operation of a diode-ring, doubled balanced mixer used as a biphas modulator. The saturation and leakage effects are not modeled, but there is a highpass filter used to simulate the transformer coupling of the mixer. Figures 13 through 16 demonstrate the model and simulation results.

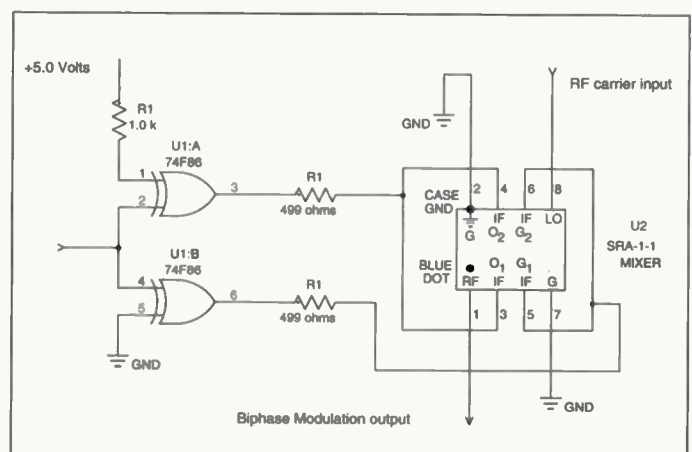


Figure 12. Schematic of a biphasic modulator.

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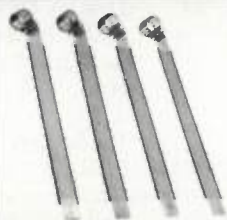
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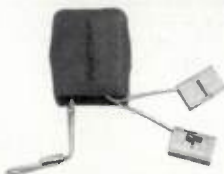
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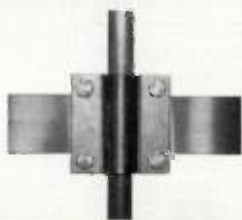
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Conclusion

SystemView is available now for \$2450.00. Optional add-on packages include a user C-code interface, Comm library, Logic library, available now. A DSP library, and a RF library are

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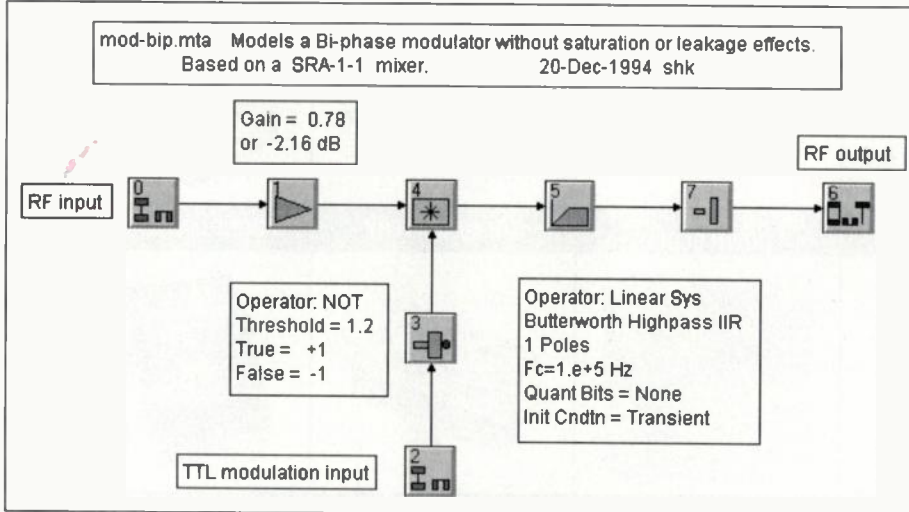


Figure 13. Model of a biphas modulator with RF and TTL inputs.

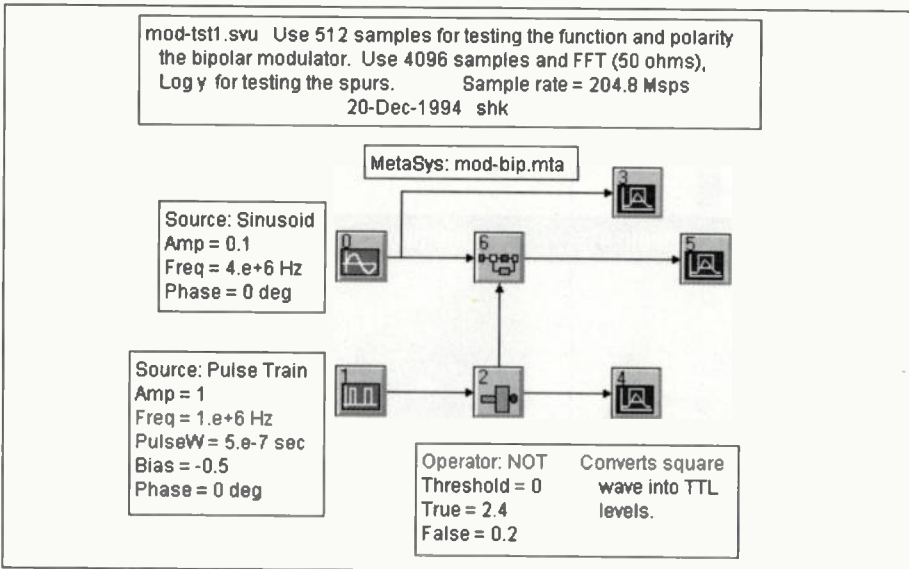


Figure 14. Test circuit for the biphas modulator.

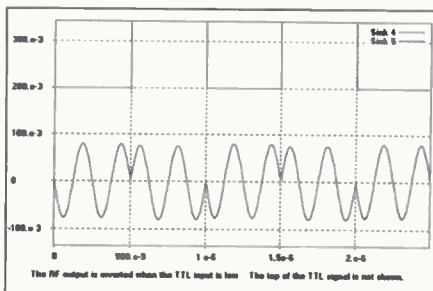


Figure 15. RF input: 4 MHz; TTL input: 1 MHz.

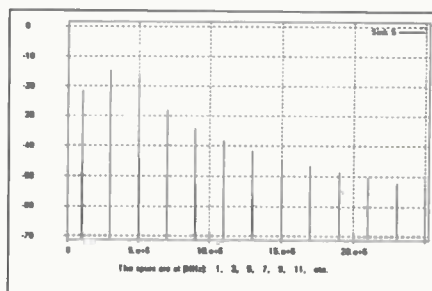


Figure 16. FFT plot of 4096 samples.

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Page numbers in () refer to the third edition of Dixon. This list contains all references for Parts 1 and 2.

Part 2 of this article will appear in the May issue of *RF Design*.

About the Author



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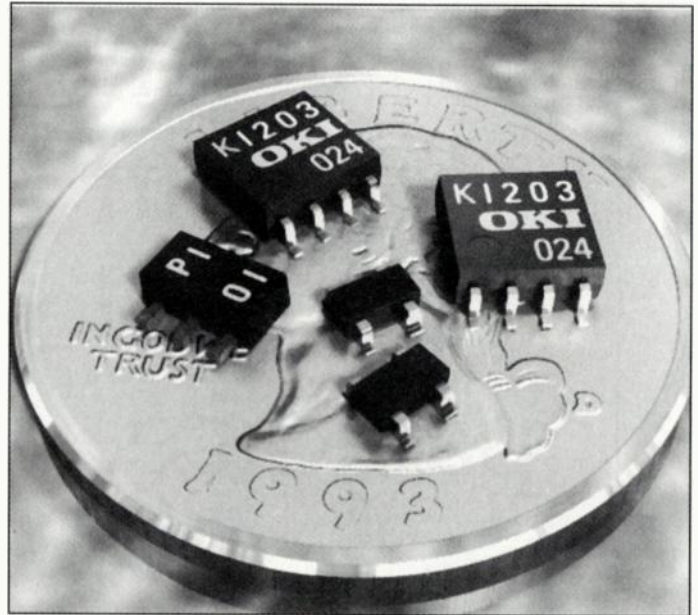
RF products

3V RF Devices

Low-power gallium arsenide (GaAs) microwave RF products operating from 850 MHz to more than 2 GHz for cellular, voice and data communications were announced by Oki Semiconductor. The high efficiency, 3V devices have been optimized and characterized to provide the necessary functions for portable UHF transceiver circuits. Among OKI's new low-power GaAs RF building blocks are — low-noise, high-gain, small signal MMIC amplifiers, (KGF1175B, KGF1521, KGF1522); high gain, high isolation broadband amplifiers, (KGF1191); mixers featuring high conversion gain and high local isolation, (KGF1203, KGF1531); digital 1/128-129 prescaler for low-power PLLs,

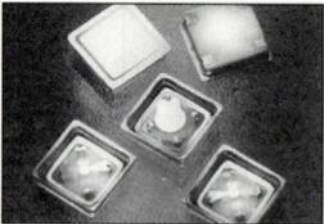
(KGL2135); GaAs MMIC drivers and power FETs for maximum legal power for high efficiency UHF transmitters, (KGF1256B, KGF1284, KGF1313). The devices are available in the four-pin, low cost SOT 143, miniature eight-pin small outline surface mount, and SC62 plastic power packages. The devices can also be specified in low-loss, high-performance ceramic packages. Prices range from \$1.76 per unit in quantities of 100 to 1,000 for the KGF1175 low-noise, high gain front end amplifier, to \$6.24 per unit in quantities of 100 to 1,000 for the KGF1313 plastic FET, with 27 dBm output power at 1.9 GHz and 3 V.

OKI Semiconductor
INFO/CARD #250



100 MHz SMT Xtals

Available in a new low profile, fully surface mountable package, the CMX 5000 series of quartz



crystal resonators from C-MAC Quartz Crystals covers the frequency range of 10 to 250 MHz. Fundamental frequencies up to 50 MHz can be achieved with conventional lapping and polishing techniques, while fundamental frequencies up to 100 MHz can be achieved with C-MAC's own plasma etching technology. Achievable frequency stabilities range from ± 5 ppm from -10 to $+60$ °C to ± 50 ppm from -55 to $+105$ °C. Long term aging is typically 1 ppm per year. The surface mount packages are based on a ceramic enclosure design with a maximum overall height of 2.7 mm. Internally, the crystals are mounted at four points rather than two, resulting in typical frequency shifts of less than 1 ppm after bump, shock and vibrations tests to IEC 68 standards.

C-MAC Quartz Crystal Ltd.
INFO/CARD #249

Ceramic Resonator Filters

RLC Electronics introduces the Model CDF-900-10-4, band-pass filter. This unit is a ceramic dielectric resonator filter designed to achieve a low loss, temperature stable, bandpass response at 900 MHz. Model CDF-900-10-4 features a four-section response with a 3 dB bandwidth of 10 MHz and a 20 dB bandwidth of 20 MHz maximum. The insertion loss at 900 MHz is less than 6 dB and the passband VSWR is 1.5:1 maximum. The unit is specified to operate over the full -54 to $+95$ °C military temperature range. Other ceramic dielectric resonator filters are available with 3 dB bandwidths of 0.7 to 12 percent of the center frequency.



Standard products feature two to seven sections and are available in SMA, connectorized, surface mount, or PC mount configurations. Prototype units are available from two to four weeks and prices start at \$195.00 in unit quantities.

RLC Electronics, Inc.
INFO/CARD #248

Spectrum Analyzer

Wandel & Goltermann's latest addition to their SNA line of spectrum analyzers provides coaxial measurements from 20 Hz to 26.5 GHz. The SNA-33 is based on fundamental frequency mixing using a unique local oscillator frequency to achieve a 90 dB dynamic range with minimal filtering. The swept phase lock synthesizer design transfers the frequency accuracy of an internal crystal reference oscillator to all frequencies in a



swept measurement. The instrument offers an extremely low noise floor, typically better than -115 dBm across the full measurement range and typically better than -127 dBm at frequencies below 3.2 GHz. The SNA-33 features frequency accuracy of 0.01 ppm. The instrument covers an amplitude measurement range of -115 to $+30$ dBm with a 0.1 dB resolution and ± 2.7 dB amplitude accuracy. Price for the SNA-33 is approximately \$50,000.

Wandel & Goltermann, Inc.
INFO/CARD #247

Through-Holes for Prototypes

LPKF CAD/CAM Systems has introduced the AutoContac System for automating through-



hole plating of circuit boards. This system dispenses an epoxy conductive ink with a filler of solderable copper grains (patent pending). The ink that LPKF is using has been formulated to maintain solderability. The AutoContac System is auxiliary to and may be mounted on all models of LPKF systems and is field installable by the user. The system consists of a dispenser, controller and interface connector. The whole plating process, including a 130 °C cure cycle, takes less than one hour. The resulting plated hole has a sidewall thickness of about 4 mils and through hole resistance of 200 milliohms (28 mil hole) to 50 milliohms (52 mil hole). The AutoContac System is \$2950. Ink costs about eight cents per hole and comes in a premix cartridge.

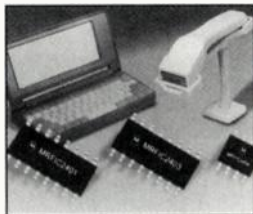
LPKF CAD/CAM Systems Inc.
INFO/CARD #246

SEMICONDUCTORS

2.4 GHz Chip Set

Motorola has introduced a three chip set of GaAs integrated circuits designed as an integrated solution for the wireless data market in the 2.4 to 2.5 GHz ISM band. The MRFIC2401 downconverter consists of a low noise amplifier and mixer with provisions for off-chip filtering after the amplifier. The MRFIC2403 power amplifier has two stages with provision for power control. The MRFIC2404 single stage exciter amplifier is intended to drive the MRFIC2403, but can be used in a variety of amplifier applications from 2 to 3 GHz. In low volume quantities, pricing for the MRFIC2401 is \$7.29, the MRFIC2403 is \$6.67, and the MRFIC2404 is \$4.03.

Motorola, Inc.
INFO/CARD #245



Phase Locked Loop IC

Hitachi America has introduced the first in its series of frequency synthesizer devices that meet the need for energy efficient cellular systems. The HD155001AT typically consumes 6 mA when operating, and 100 μ A when in standby mode. The chip requires only an external loop filter and VCO to form a complete PLL synthesizer. The HD155001AT has fast lock-up times and a prescaler with division modes programmable for 64/65 or 128/129. In quantities of 10,000, the HD155001AT is \$2.95.

Hitachi America, Ltd.
Semiconductor & I.C. Div.
INFO/CARD #244

CATV Converters

Anadigics' 860 MHz GaAs chip-set offers an integrated alternative to the use of traditional discrete components for CATV applications. The ACU50750 upconverter IC accepts TV signals in the 50 to 860 MHz range and converts them to an IF in the 900 to 1200 MHz range. The ACD0900 downconverts an IF the 900 to 1200 MHz range to a second IF between 35 and 150 MHz. Both ICs feature an on-board oscillator /phase splitter and double balanced mixer. The chip set is priced starting at \$5.20 for quantities of 10,000 per month.

Anadigics
INFO/CARD #243

GSM Baseband Chips

A kernel processor and vocoder which comprise a two-chip solution for GSM baseband signal processing has been introduced by VLSI Technology's Wireless Products Division. The VP22002 kernel processor

integrates Type Approved GSM functional blocks, such as channel coder, equalizer, GMSK modulator and timing generator.

VLSI Technology, Inc.
INFO/CARD #242

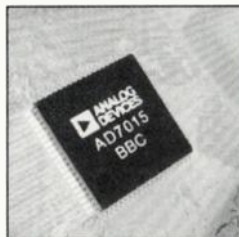
GaAs MMIC Amplifier

The AH-B102D-3 series of low cost, monolithic HBT GaAs MMIC amplifiers is part of the family of cascadable HBT Darlington amplifiers from FEI Communications. The AH-B102D-3 features typical gain of 22 dB, RF frequency of DC to 3 GHz, < 5 dB noise figure, 1 dB compression point of > 10 dBm, and typical input/output VSWR of 2.0:1 (below 1 GHz). Price starts at \$3.25 in 100k quantities.

FEI Communications, Inc.
INFO/CARD #241

Wireless Baseband Chip

The AD7015 integrates the multitude of mixed-signal components required for a GSM, PCS, or PCN handset into one IC. The chip operates from a 3V supply and includes eleven converters, together with digital filters, modulator, amplifiers, multiplexers, references, and logic to implement all the complex signal conversion required



in a wireless handset. The AD7015 contains three independent sections: a voice-band codec, base-band codec, and an auxiliary stage which performs control and monitoring functions.

The device is sampling now and will be available in volume in mid-1995. It will be priced at \$13 in OEM quantities.

Analog Devices, Inc.
INFO/CARD #240

AMPLIFIERS

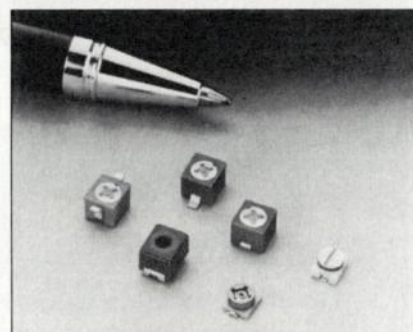
Power Amplifier Hybrid

Microwave Technology has announced the MPS-093011, a surface mountable, linear power amplifier module for base station/infrastructure applications. The hybrid amplifier operates from 800 to 1000 MHz, has a P_{1dB} of +30 dBm, IP3 of +45 dBm, and a small signal gain of 16 dB. It operates from a single +7.5 V supply and typically consumes 400 mA. The amplifier has 50 Ω input/output impedances.

Microwave Technology
INFO/CARD #239

Improved Wideband Amp

IFI has completed a major redesign of the M410, 2200 W tube amplifier, and now offers the half-sized, more power efficient TCCX 2200. The single cabinet, user serviceable amplifier offers 2200 W linear CW power at



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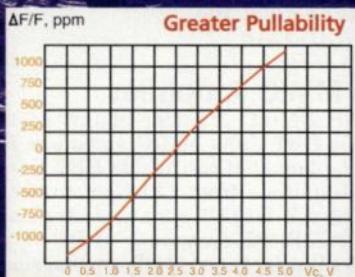
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Instruments for Industry
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Low Noise, Digital Data Amp

Model HA-103-35 from Hypres is a high gain, wideband amplifier module with a gain of 60 dB into a 50 Ω load. The amplifier gain is flat to 0.2 dB from DC to 20 MHz and has a -3 dB point at 35 MHz. The amplifier has an input noise level of 1.4 nV/√Hz. Model HA-103-35 is optimized for the amplification of low level digital data and has a -1 dB compression point at 1.8 mV. Pricing is \$595 for quantities less than 10.

Hypres, Inc.
INFO/CARD #237

Linear Cellular & PCN Amps

Phoenix Microwave has introduced a line of ultra linear modular amplifiers for use in cellular and PCN/PCS systems. PA1062 covers the 850 to 925 MHz frequency range with +34 dBm output power, 27 dB gain, 1.8 dB noise figure and 3rd order intercept of +49 dBm. PA1074 covers the 1800 to



2000 MHz PCN band with performance levels slightly reduced as compared with the PA1062 module. The entire family is packed in miniature flanged plastic packages for surface mount assembly.

Phoenix Microwave Corp.
INFO/CARD #236

Ultra-Broadband, 30W

A solid state RF power amplifier from Amplifier Research delivers 30 W of CW power across a frequency range of 25 MHz to 1 GHz. Model 30W1000M7 is specially designed for susceptibility and other RF testing applications requiring level, consistent output power over the test bandwidth. At maximum gain setting, the amplifier provides 30 W minimum over the entire bandwidth, even in cases of infinite VSWR or open or shorted terminals. Gain is controllable over a 10 dB range. Price is \$9,000.

Amplifier Research
INFO/CARD #235

TOOLS, MATERIALS & MANUFACTURING

RF Prototyping

ProtoCell is a modular prototyping system designed for rapid fabrication and char-

acterization of RF signal processing circuits. The patent pending design consists of a rigid aluminum frame with 36 cell positions to accommodate circuit boards which fit a wide variety of circuit functions and component packages.



Each cell contains a circuit function such as a mixer, amplifier, filter, etc. These individual cells attach to the frame and the cells are then joined by a

simple solder bridge. The RF Frame sells for \$250.00, individual boards vary in price, but sell for an average of \$12.00 each.

Innovative Technology
INFO/CARD #234

Microwave Substrates

New CTLE (Controlled Low Thermal Expansion) laminated microwave substrate materials using ARLON's proprietary formulation have reduced Z-direction thermal expansion. The materials also exhibit high thermal conductivity and minimal change in ϵ_r caused by PTFE's second order phase transition at 19°C. The dielectric constant of CLTE materials is 2.94. The laminates are supplied with 0.5, 1, or 2 ounce electrodeposited copper on both sides; other platings are available.

ARLON
INFO/CARD #233

Circuit Board Material

Rogers introduces RO4003™ high frequency circuit board material. The material, which is based on a ceramic filled thermoset dielectric, can be fabricated by nearly any board shop using nearly standard material fabrication processes for epoxy glass boards. RO4003 offers a κ' of 3.38, dissipation factor of 0.0022 at 10 GHz, and a thermal coefficient of expansion in the Z-axis of < 35 ppm. The material is priced at \$8.00/ft² for 0.020" laminates in high volume.

Rogers Corp.
INFO/CARD #232

TEST EQUIPMENT

CDMA Signal Source

The CDMA Coder for the SMHU.58 signal generator produces CDMA signals with a chip rate of 1.2288 MHz. Up to two channels of a CDMA-coded PCS base station can be emulated, with channel 1 as the pilot channel and channel 2 configurable for pilot, sync, paging, or traffic channel coding. Channel coding is according to IS-95, and calculation of CRC, the encoder tail, convolution encoding, block interleaving, and Walsh coding are included. The CDMA Coder is priced in the U.S. at \$6,340, or it

may be purchased installed in a SMHU.58 signal generator for \$49,240.

Tektronix, Inc.
INFO/CARD #231

Channel Emulator

The TAS 4500 FLEX™ emulates the delay, fast and slow fading and path loss characteristics of RF mobile communication channels. The emulator offers cartridge style local oscillators, RF front-end modules, and IF signal processing modules.

The 4500 FLEX also features a built-in RF attenuator. The list price for the 4500 FLEX base unit is \$40,000
Telecom Analysis Systems, Inc.
INFO/CARD #230

DDS Function Generator

Pragmatic Instruments announces its Model 1404A, a direct digital synthesized 20 MHz function generator. The generator covers 100 mHz to 20 MHz with 10 mHz resolution and offers frequency sweeps over

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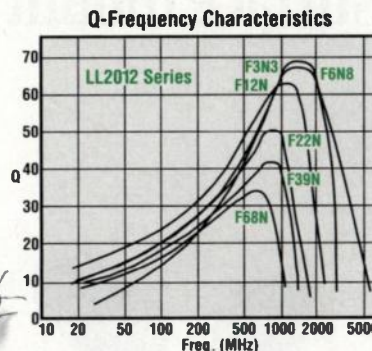
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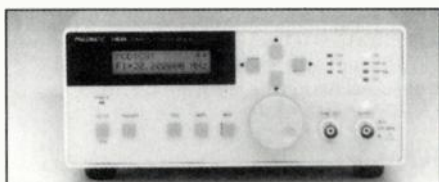
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this entire range with linear or logarithmic profiles going both up and down in frequency. The user has direct access to modulation parameters and can generate a variety of modulation types. Model 1404A has a list price of \$1,295.

Pragmatic Instruments, Inc.
INFO/CARD #229

Xtal Product Analyzer

Hewlett-Packard's 300 MHz network analyzers focus on the final testing and plating process in the production test of crystal filters and resonators. The analyzer covers 10 kHz to 300 MHz with 1 mHz resolution. Output power is -9 to +11 dBm, with an optional -48 to +22 dBm range. IF bandwidths from 10 Hz to 30 kHz are selectable. The HP E5100A is priced starting from \$15,000.

Hewlett-Packard Co.
INFO/CARD #228

VXIbus Synthesizers

Communications Techniques has introduced a narrowband VXIbus synthesizer designed specifically for cellular and WLAN test applications. The single slot, C size module operates at frequencies from 820 to 960 MHz and 2.4 to 2.5 GHz. Typical phase noise for a 900 MHz unit with 10 kHz resolution is -115 dBc at 1 kHz. Spurious levels are less than -80 dBc. The products are in full compliance with VXIbus specification rev. 1.4 and are SCPI compatible.

Communications Techniques, Inc.
INFO/CARD #227

SIGNAL PROCESSING COMPONENTS

8-Way Power Divider

Pulsar model P8-06-303 is an 8-way, 0.1 to 2.0 GHz, flatpack power divider with 16 dB isolation, 4.5 dB insertion loss, VSWR <1.7:1 and phase unbalance less than 10°.

A 12-way divider is also available.
Pulsar Microwave Corp.
INFO/CARD #226

Frequency Doublers

A series of MMIC frequency doublers, introduced by Hittite, produces an output signal at twice the input signal frequency. The MMICs use no DC bias, requiring only connections to input and output ports. The HMC156 output frequency is 1.6 to 3.4 GHz, the HMC157 covers 2.4 to 5.2 GHz, and the HMC158 output frequency is 3.2 to 7.2 GHz. Conversion losses are 18, 15, and 13 dB for the 156, 157, and 158, respectively.

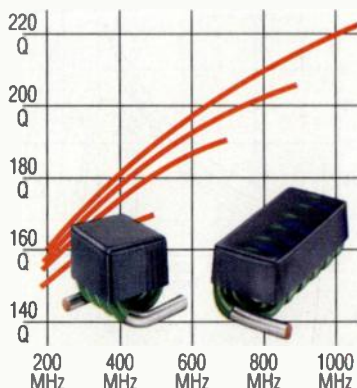
Hittite Microwave Corp.
INFO/CARD #225

Temp Compensating Attenuator

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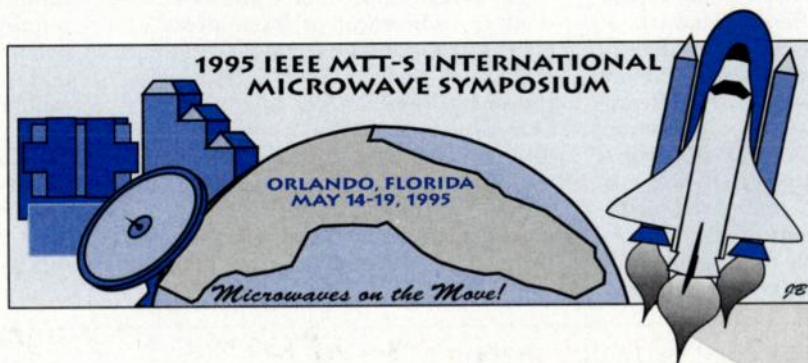
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Image Reject and Image Canceling Mixers

By Louis Pandula
VLSI Technology, Inc.

As levels of radio IC integration increase, image reject and image canceling mixers are finding renewed use in low cost, radio systems. The effective design and use of image reject and image canceling mixers requires an understanding of their basic principles of operation. This article presents the theory of operation for both mixer types by outlining the mathematics governing their operation and by encouraging an intuitive understanding of the fundamental principles embodied in the equations.

Image reject and image canceling mixers have historically found limited usefulness in consumer radio applications. These have been restricted to niche radio architectures that warranted their additional complexity. Some applications included satellite receivers with low IF frequencies for the image reject mixer and single-sideband transmitters for the image canceling mixer.

With present trends towards ever higher levels of IC integration gradually spilling over into radio system design, radio architectures that were once considered impractical for low cost consumer applications are quickly finding renewed usefulness. This is especially true as the requirements of more complex modulation formats and

duplexing techniques outstrip the capabilities of conventional, low cost, radio design approaches.

This is typified by the introduction of two chips; an RFIC as presented at ISSCC 94 by Hewlett Packard and the SA900 from Philips. Both chips provide a highly integrated radio transmit section suitable for AMPS, NADC, and CDPD applications as well as other systems requiring complex modulation capability that operate near 900 MHz. Key to the operation of both chips is an image canceling mixer used in an offset loop architecture.

Certainly even more highly integrated radio chips will follow as semiconductor manufacturers learn to exploit the capabilities of these two versatile building blocks.

Image Reject Mixer

The image reject mixer allows a heterodyne radio receiver to suppress RF energy at the image frequency without the use of an RF filter. Historically, the image reject mixer has been used in receiver applications with a very low first IF where the RF front-end filter provides little attenuation at the image frequency. This is typical of satellite receivers and a number of industrial, scientific, and medical (ISM) receiver designs.

In conventional heterodyne receivers, a bandpass filter provides some level of image suppression by attenuating all frequencies outside the band of interest. As an example, a heterodyne radio designed to receive signals in the 902 to 928 MHz ISM band might use a first intermediate frequency (IF) at 70 MHz. With a low-side local oscillator (LO) at 832 to 858 MHz, this places the image frequency band at 762 to 788 MHz, or 140 MHz away from the desired frequency. As illustrated in Figure 1, a bandpass filter centered at 915 MHz can provide some level of image suppression, typically 20 to 40 dB.

An image reject mixer also provides image frequency suppression, however, it uses phase cancellation rather than frequency selective attenuation to achieve it. In practice, however, it may be necessary to use a bandpass filter in conjunction with the image reject mixer to prevent high power, out of band signals from overloading the RF front end and generating spurious products that could fall into the IF passband.

When used together, the combination of bandpass filter and image reject mixer provides superior image suppression over either technique used independently. In addition, the image reject mixer achieves a lower receiver noise figure by suppressing

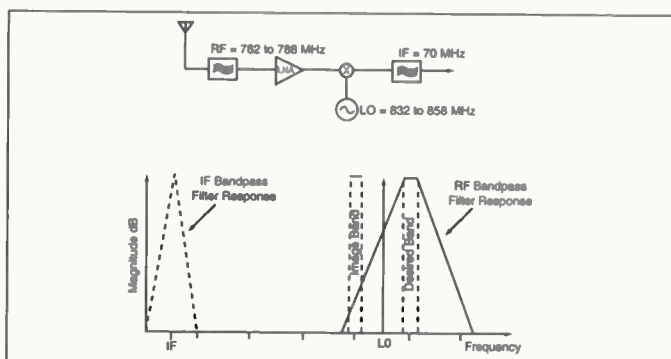


Figure 1. Traditional heterodyne radio receiver and frequency plan.

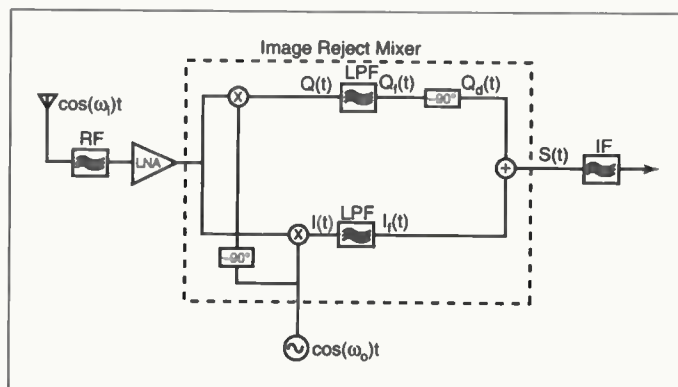


Figure 2. Heterodyne receiver with image reject mixer.

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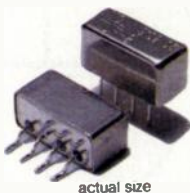
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SPECIFICATIONS

Model	LO Power (dBm)	Freq. LO/RF (MHz)	■ Conv. Loss (dB) \bar{X} δ	Isol. L-R (dB)	Price.\$ Ea. 10 qty
TUF-3	7	0.15-400	4.98 0.34	46	5.95
TUF-3LH	10		4.8 0.37	51	7.95
TUF-3MH	13		5.0 0.33	46	8.95
TUF-3H	17		5.0 0.33	50	10.95
TUF-1	7	2-600	5.82 0.19	42	3.95
TUF-1LH	10		6.0 0.17	50	5.95
TUF-1MH	13		6.3 0.12	50	6.95
TUF-1H	17		5.9 0.18	50	8.95
TUF-2	7	50-1000	5.73 0.30	47	4.95
TUF-2LH	10		5.2 0.3	44	6.95
TUF-2MH	13		6.0 0.25	47	7.95
TUF-2H	17		6.2 0.22	47	9.95
TUF-5	7	20-1500	6.58 0.40	42	8.95
TUF-5LH	10		6.9 0.27	42	10.95
TUF-5MH	13		7.0 0.25	41	11.95
TUF-5H	17		7.5 0.17	50	13.95
TUF-860	7	860-1050	6.2 0.37	35	8.95
TUF-860LH	10		6.3 0.27	35	10.95
TUF-860MH	13		6.8 0.32	35	11.95
TUF-860H	17		6.8 0.31	38	13.95
TUF-11A	7	1400-1900	6.83 0.30	33	14.95
TUF-11ALH	10		7.0 0.20	36	16.95
TUF-11AMH	13		7.4 0.20	33	17.95
TUF-11AH	17		7.3 0.28	35	19.95

*To specify surface-mount models, add SM after P/N shown.

■ \bar{X} = Average conversion loss at upper end of midband ($f_u/2$)
 δ = Sigma or standard deviation

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noise components at the image frequency. A conventional heterodyne receiver would require a second band-pass filter following the low noise amplifier (LNA) to achieve an equivalent noise figure.

Figure 2 shows that the image reject mixer consists of a pair of mixers driven from a quadrature LO source, a pair of lowpass filters, a 90° phase shifter, and a power combiner.

To understand the reject mixer's operation it is important to examine how both the desired frequency and the image frequency are affected by the quadrature down-conversion process. For simplicity ω is used throughout to represent $2\pi f$.

The in-phase mixer product equals:

$$I(t) = \cos(\omega_i t) \cdot \cos(\omega_0 t) \quad (1)$$

and this can be rewritten as:

$$I(t) = 1/2\cos(\omega_i + \omega_0)t + 1/2\cos(\omega_i - \omega_0)t \quad (2)$$

The quadrature mixer product equals:

$$Q(t) = \cos(\omega_i t) \cdot \cos(\omega_0 t - 90^\circ) \quad (3)$$

and this can be rewritten as:

$$Q(t) = 1/2\cos(\omega_i t + \omega_0 t - 90^\circ) + 1/2\cos(\omega_i t - \omega_0 t + 90^\circ) \quad (4)$$

After lowpass filtering only the difference frequency terms remain and the outputs become:

$$I_f(t) = 1/2\cos(\omega_i - \omega_0)t \quad (5)$$

$$Q_f(t) = 1/2\cos((\omega_i - \omega_0)t + 90^\circ) \quad (6)$$

Now let's assume a low side LO and let's see what happens when we run the two signal frequencies through equations 5 and 6 above. For the desired signal, ω_i is greater than ω_0 and the difference frequency $(\omega_i - \omega_0)$ yields a positive value. Substituting ω_d for $|\omega_i - \omega_0|$ the previous equations can be rewritten as follows:

$$I_f(t) = 1/2\cos(\omega_d t) \quad (7)$$

$$Q_f(t) = 1/2\cos(\omega_d t + 90^\circ) \quad (8)$$

The quadrature path can be rewritten as:

$$Q_d(t) = -1/2\sin(\omega_d t) \quad (9)$$

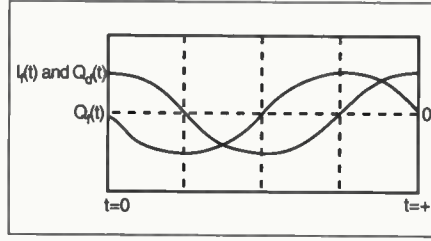


Figure 3a. Relative waveforms for $I_f(t)$, $Q_f(t)$, and $Q_d(t)$ for $\omega_i > \omega_0$.

and after applying an additional -90° phase shift, its output becomes:

$$Q_d(t) = -1/2\sin(\omega_d t - 90^\circ) \quad (10)$$

which simplifies to:

$$Q_d(t) = 1/2\cos(\omega_d t) \quad (11)$$

After summing the in-phase and quadrature signals in equations 7 and 11, the composite signal becomes:

$$S(t) = 1/2\cos(\omega_d t) + 1/2\cos(\omega_d t) \quad (12)$$

which reduces to:

$$S(t) = \cos(\omega_d t) \quad (13)$$

From the results above, it is clear that the received signal combined constructively after passing through the image reject mixer and was simply translated to an IF frequency corresponding to ω_d . This can be seen in Figure 3a with the quadrature signal leading the in-phase signal by 90° and the phase shifted quadrature signal being superimposed on top of the in-phase signal.

Now let's examine how the image reject mixer processes a signal at the image frequency. In this case ω_i is less than ω_0 and the difference frequency $(\omega_i - \omega_0)$ yields a negative value. Again

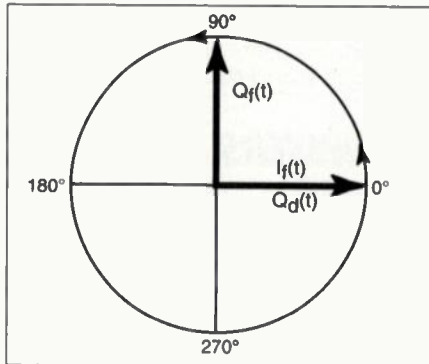


Figure 4a. Vector diagram of $I_f(t)$, $Q_f(t)$, and $Q_d(t)$ for $\omega_i > \omega_0$ at $t=0$.

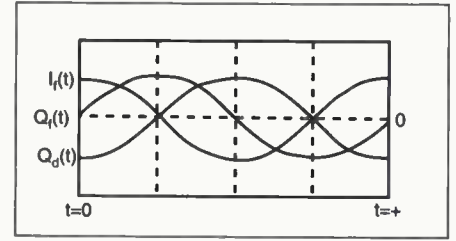


Figure 3b. Relative waveforms for $I_f(t)$, $Q_f(t)$, and $Q_d(t)$ for $\omega_i < \omega_0$.

substituting ω_d for $ABS(\omega_i - \omega_0)$ the outputs after filtering become:

$$I_f(t) = 1/2\cos(-\omega_d t) = 1/2\cos(\omega_d t) \quad (14)$$

$$Q_f(t) = 1/2\cos(-\omega_d t + 90^\circ) = 1/2\sin(\omega_d t) \quad (15)$$

Again applying a -90° phase shift to the quadrature channel:

$$Q_d(t) = 1/2\sin((\omega_d t - 90^\circ) \quad (16)$$

which simplifies to:

$$Q_d(t) = -1/2\cos(\omega_d t) \quad (17)$$

Finally, summing the in-phase and quadrature signals, the composite signal becomes:

$$S(t) = 1/2\cos(\omega_d t) + -1/2\cos(\omega_d t) \quad (18)$$

which reduces to:

$$S(t) = 0 \quad (19)$$

As we expected, for the case of the image frequency, the signals through the image reject mixer combined destructively causing complete cancellation of the signal. This can be seen in Figure 3b with the quadrature signal now lagging the in-phase signal and the phase shifted quadrature sig-

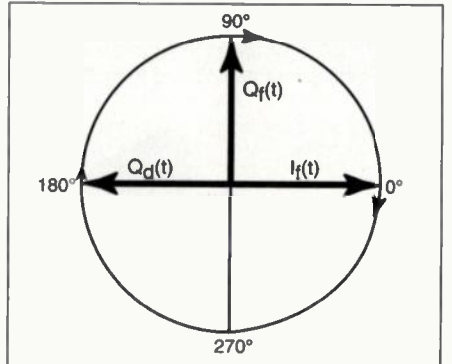


Figure 4b. Vector diagram of $I_f(t)$, $Q_f(t)$, and $Q_d(t)$ for $\omega_i < \omega_0$ at $t=0$.

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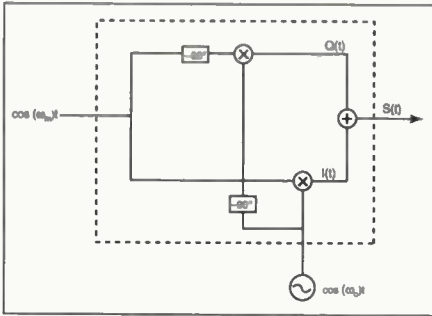


Figure 5 Image canceling mixer for single sideband generation.

nal taking on the exact opposite phase as the in-phase signal.

For a more intuitive understanding, let's examine Figures 4a and 4b for a graphical representation of the equations above. The instantaneous phase of the signals at the two filtered outputs $I_f(t)$ and $Q_f(t)$ is represented by a vector rotating around the origin at a rate ω_d . The vector's direction of rotation depends on the sign of $(\omega_i - \omega_0)$ with positive values causing a counter clockwise rotation and negative values causing a clockwise rotation. The extension of the vector onto the horizontal axis is equal to the real, instantaneous magnitude of the signal as would be observed on an oscilloscope.

Figure 4a illustrates the vector positions at $t = 0$ for the desired signal case, when ω_i is greater than ω_0 . From equation 7, the $I_f(t)$ output is shown starting at 0° and having a counter clockwise rotation. From equation 8, the $Q_f(t)$ output is shown starting from $+90^\circ$ and also having a counter clockwise rotation. Now if we apply a -90° phase shift to the $Q_f(t)$ vector, this is graphically equivalent to retarding the $Q_f(t)$ vector 90° clockwise from its starting position. As a result the $Q_d(t)$ vector falls on top of the $I_f(t)$ vector and the two sum constructively.

Figure 4b illustrates the image signal case, when ω_i is less than ω_0 . From equations 14 and 15, the initial vector positions at $t = 0$ are the same as above. However, now their rotation is in the clockwise direction. This is a very significant difference. Again applying a -90° phase shift to the $Q_f(t)$ signal causes it to be retarded in the counter clockwise direction to a position at 180° . As a result, for the image frequency case, the $Q_d(t)$ vector is 180° out of phase with the $I_f(t)$ vector and the two cancel.

Of course amplitude and phase balance between the I and Q paths ultimately determine the amount of image

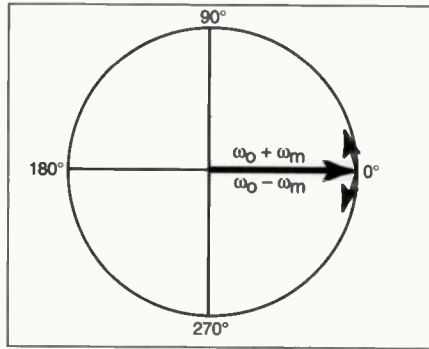


Figure 6a. Vector diagram for I-channel sum and difference terms at $t = 0$, shown relative to ω_0 .

frequency rejection that is achievable, but 20 to 40 dB is feasible with practical devices. With DSP implementations on the other hand, the performance is only limited by the available arithmetic precision.

Image Canceling Mixer

The image canceling mixer allows a signal to be translated from one frequency to another in a manner that generates a single-sideband signal. In other words, either the sum or difference frequency is generated but not both. Traditionally, this mixer has been used for single-sideband transmission but recently its properties are being harnessed in radio architectures that use an IF frequency in the transmit path. Here the image canceling mixer allows the IF signal to be translated up to the carrier band without the need for highly selective filters to suppress the unwanted image.

The image canceling mixer, as illustrated in Figure 5, uses a structure almost identical to the image reject mixer except for the position of the phase shifter in the signal path is moved to the input side. Again, the heart of the structure is a quadrature driven mixer pair with a combining network at its output.

To understand the operation of the image canceling mixer it is important to examine what happens to the input signal in both the in-phase and quadrature signal paths. The in-phase mixer product equals:

$$I(t) = \cos(\omega_0 t) \cdot \cos(\omega_m t) \quad (20)$$

and this can be rewritten as:

$$I(t) = 1/2 \cos(\omega_0 + \omega_m)t + 1/2 \cos(\omega_0 - \omega_m)t \quad (21)$$

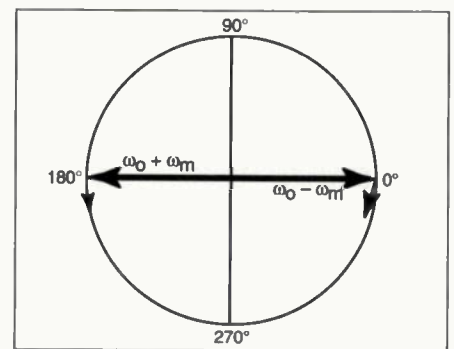


Figure 6b. Vector diagram for Q-channel sum and difference terms at $t = 0$, shown relative to ω_0 .

The quadrature mixer product equals:

$$Q(t) = \cos(\omega_0 t - 90^\circ) \cdot \cos(\omega_m t - 90^\circ) \quad (22)$$

and this can be rewritten as:

$$Q(t) = 1/2 \cos(\omega_0 t - 90^\circ + \omega_m t - 90^\circ) + 1/2 \cos(\omega_0 t - 90^\circ - \omega_m t + 90^\circ) \quad (23)$$

Simplifying the quadrature product above yields:

$$Q(t) = 1/2 \cos((\omega_0 + \omega_m)t - 180^\circ) + 1/2 \cos(\omega_0 - \omega_m)t \quad (24)$$

Rewriting the first term in the above equation gives:

$$Q(t) = -1/2 \cos(\omega_0 + \omega_m)t + 1/2 \cos(\omega_0 - \omega_m)t \quad (25)$$

Summing the in-phase and quadrature signals from 9 and 11 above:

$$S(t) = 1/2 \cos(\omega_0 + \omega_m)t + 1/2 \cos(\omega_0 - \omega_m)t - 1/2 \cos(\omega_0 + \omega_m)t + 1/2 \cos(\omega_0 - \omega_m)t \quad (26)$$

Notice that the two sum frequency terms cancel leaving:

$$S(t) = 1/2 \cos(\omega_0 - \omega_m)t + 1/2 \cos(\omega_0 - \omega_m)t \quad (27)$$

Finally, simplifying leaves only the lower sideband:

$$S(t) = \cos(\omega_0 - \omega_m)t \quad (28)$$

Figures 6a and 6b provide a pictorial view of the equations above. Again, the instantaneous phase of each signal term is represented by a rotating vec-

tor, but now instead of using zero Hertz as a reference, the vector rotation is shown relative to ω_0 .

When using this notation it can be shown that the instantaneous magnitude of the composite signal represented in the vector diagram is equal to the sum of the real components of the individual vector magnitudes calculated with respect to an axis rotating clockwise at a rate ω_0 .

Figure 6a uses this notation to illustrate the two terms in equation 21 for the I channel. The difference frequency term is shown as a vector rotating in the clockwise direction while the sum frequency term is shown rotating in the counter clockwise direction.

As expected, if we were to observe this signal on an oscilloscope, it would appear as the carrier signal at ω_0 modulated by a sinusoidal signal at a rate equal to ω_m with a modulation index of 100 percent.

Similarly, Figure 6b shows the two terms for the Q channel output derived in equation 25. Again the sum and difference components are represented by counter-rotating vectors, but now their phase is shifted by 180° relative to each other.

Observing this second composite signal on an oscilloscope we would see that both the carrier phase and envelope phase would exhibit a 90° phase lag relative to the I channel signal.

Finally by superimposing Figure 6a on top of Figure 6b, it is apparent that the I and Q channel sum terms are opposite in phase and cancel while the difference terms add constructively. In essence, the upper sideband is canceled while the lower sideband is passed unaltered as predicted by equation 26.

Amplitude and phase unbalance between the I and Q paths limit the amount of image frequency cancellation that is achievable in practice, with devices achieving 20 to 50 dB available in both hybrid and monolithic form.

Conclusion

The theory of operation for both image reject and image canceling mix-

ers has been provided. This included a derivation of the mathematics governing their operation and an explanation of the fundamental principles embodied in the equations. By encouraging both a theoretical and an intuitive understanding this article hopes to promote the effective design and use of these two key radio building blocks in the next generation of low cost, radio systems. **RF**

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About the Author



Louis Pandula is a Staff Design Engineer with VLSI Technology's Wireless Product Division. He immigrated to the United States

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Linear Circuit Analysis Program Uses Two-Port Method

By Dale Henkes
Phillips Consumer Electronics Co.

This program, called **LINC** (for **Linear Circuit Analysis**), is a circuit simulation program written for the Microsoft Windows™ graphical operating environment. The program uses linear two-port analysis techniques to evaluate electrical circuits described in a well established two-port circuit language. A rich set of analysis tools and circuit performance indicators are provided. Additionally, the program offers many ways of displaying output, including Smith Charts, rectangular graphs, tabulated numeric data as well as printed output of the above.

LINC version 1.1 features a built-in full-screen text editor for generating circuit files, S parameter data and circuit file importation and exportation, and a set of 18 analysis responses and performance indicators. Among those analyzed responses are: magnitude (dB or linear) and phase of forward gain (S_{21}), reverse gain (S_{12}), input reflection coefficient (S_{11}), and output reflection coefficient (S_{22}). The program also calculates group delay, maximum stable gain (MSG), stability indicators K and Delta, input impedance (R_{in} and X_{in}), output impedance (R_{out} and X_{out}), and input and output VSWR.

LINC also produces Smith chart displays of S_{11} and S_{22} , and stability circles on the input and output reflection coefficient planes.

How The Program Works

This program is based on the principle that if the overall circuit and its components are linear and time invariant, then each component can be treated as a linear two-port "black box". The overall circuit can then be constructed from the individual component two-ports. This requires that all components in the circuit be thought of as two-ports and that their interconnections be expressed as connections of two-ports. This is done by

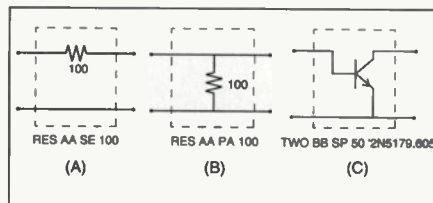


Figure 1. Examples of simple two-port components.

creating a circuit file that identifies each circuit element as a two-port along with its connect code.

Circuit Files

A circuit file is a text description of an electrical circuit in a format that the circuit analysis program can interpret and analyze. The circuit description format used in this program is similar to that used by several commercially available circuit analysis and simulation programs of the past and present. In fact, circuit files can be written that will run interchangeably without modification on this program and programs such as Eagleware's STAR and SuperStar Professional™. This program was designed to have a high degree of compatibility with the earlier STAR and SuperStar programs by Eagleware. Circuit files that have the "CIRCUIT OLD" and "WINDOW" block descriptors will also run on the latest version of SuperStar Professional.

The body of the circuit file is a series of three letter mnemonics or codes that represent all the circuit elements (electrical components), operations to be performed on the circuit elements, and connect codes describing how the components are connected. There is an OUTPUT statement that specifies the resultant two-port circuit to be analyzed and the system impedance or terminations at the source and load. Finally, there is a sweep specification that defines the frequencies over which the circuit will be analyzed.

Table 1 contains the list of circuit

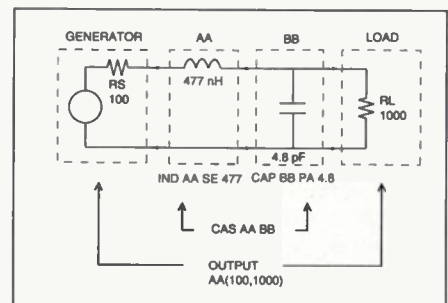


Figure 2. Example of cascaded two-ports

components supported by this program.

Circuit Components As Two-Ports

Two terminal components that are in series with the signal flow path are given an SE designation. Figure 1A shows a two-port constructed from a 100 ohm resistor in a series orientation. Below the figure is the circuit file code required to describe it to the program. The first three letters indicate the type of part (resistor, capacitor, inductor etc.). The next two letters can be thought of as an identifier tag, uniquely labeling it from other similar parts.

Two terminal components that are in parallel with the signal flow path are given a PA designation. Figure 1B shows a two-port constructed from the same 100 ohm resistor in a parallel (PA) orientation. Again, the circuit codes are shown below the figure.

Circuit components that have more than two terminals are not given an SE or PA designation. The transistor in Figure 1C is an example. The transformer, transmission line and transmission line stub section, are other examples of components that are not designated as SE or PA.

The "TWO" code in Figure 1C directs the program to import S parameter data from an external file and store it at the location BB. In Figure 1C the two-port data is contained in a file named

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CAP	ideal and finite Q capacitor
IND	ideal and finite Q inductor
OST	open circuit stub transmission line section
PLC	ideal and finite Q parallel LC resonator
PRC	parallel resistor and capacitor combination
PRL	parallel resistor and inductor combination
PRX	parallel resistor, inductor and capacitor
RES	resistor
SLC	ideal and finite Q series LC resonator
SRC	series resistor and capacitor combination
SRL	series resistor and inductor combination
SRX	series resistor, inductor and capacitor
SST	short circuit stub transmission line section
TRF	ideal transformer
TRL	transmission line
TWO	two-port from external data file
XTL	crystal resonator

Table 1. Circuit components supported by LINC.

2n5179.605. SP 50 indicates that the data is 50 ohm S parameter data in industry standard format. Imported S parameter data is usually, but not necessarily, active device data. S parameter data from any measured device (active or passive) can be imported. Also, previously computed data as output from this or other programs can constitute a valid data file. In fact, entire circuits can be saved as S parameter data and later imported as subassemblies to be reused in larger circuits.

Interconnections Of Two-Ports

The most common connection of two-ports is the cascade. The three letter mnemonic is CAS. The CAS statement connects any two two-ports in cascade. The two-ports do not have to be alphabetically consecutive. The CAX statement is similar. However, the CAX statement is used to connect two or more consecutive two-ports in cascade. The word cascade is used here because a series connection means something entirely different in two-port language. The series (SER) connection will be explained later.

The CAS operation is illustrated in Figure 2. The circuit file required to run this simulation is as follows:

```
IND AA SE 477
CAP BB PA 4.8
CAS AA BB
OUTPUT
AA(100, 1000)
SWP 75 125 51
```

Two-port AA is a 477 nH inductor in series (SE) orientation. Two-port BB is a 4.8 pF capacitor in parallel (PA) orientation. Execution of the CAS AA BB command generates a single new network that is equivalent to the cascade combination of AA and BB. The new network is stored as two-port AA and the previous contents of AA are overwritten. Two-port AA now represents the entire "L" match network between

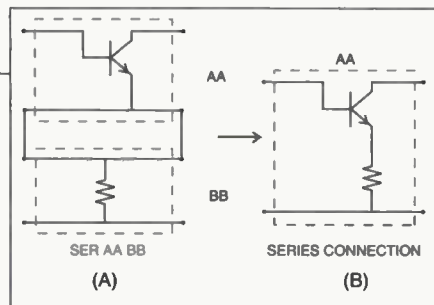


Figure 3. Example of series connection of two-ports.

the 100-ohm generator and the 1000-ohm load.

The OUTPUT statement indicates which two-port is to be analyzed. In addition, the output statement indicates the source and load terminations. The last line is the sweep statement. In this case a frequency sweep of 51 points between 75 and 125 MHz is indicated.

The SERIES (SER) connection of two-ports AA and BB is illustrated in Figure 3. This is called a series connection because both sets of input (port 1) terminals and both sets of output (port 2) terminals are connected in series. The resultant network, shown in Figure 3B, replaces two-port AA.

The PARALLEL (PAR) connection is illustrated in Figure 4. The PAR AA BB statement connects both the input and output terminals of each network in shunt (parallel). The resultant network, shown in Figure 5B, replaces two-port AA.

A broadband amplifier can now be designed using all the techniques previously discussed. Consider the schematic in Figure 5. Below is the circuit file describing this circuit:

```
CIRCUIT
'Broadband RF Amplifier
CAP AA PA 2.71 'C1
IND BB SE 6.44 'L1
TWO CC SP 50 'MRF901.615
CAP DD PA 2.71 'C2
IND EE SE 2.93 'L4
RES FF PA 5.1 'R1
SRL GG SE 259.07 6.04 'R2 L2
IND HH SE 7.98 'L3
CAS CC HH
SER CC FF
PAR CC GG
CAX AA EE
OUTPUT
AA(50)
FREQ
SWP 50 950 91
```

The first eight lines after the CIRCUIT header identify all eight components as two-ports AA through HH. The next four lines are connect codes. The first connect code (CAS CC HH) connects transistor Q1 to inductor L3 and stores it as CC. The second con-

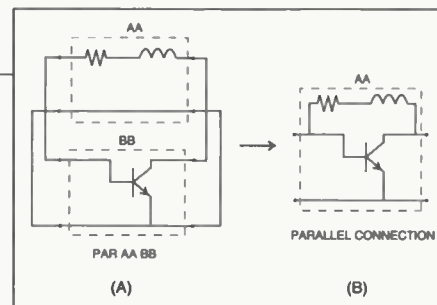


Figure 4. Example of parallel connection of two-ports.

nect code (SER CC FF) connects resistor R1 in series with the cascade combination of Q1 and L3. All three components (R1, Q1 and L3) are now stored as two-port CC. The third connect code (PAR CC GG) connects R2 and L2 in parallel with CC. The result is that two-port CC has been modified to include Q1, R1, R2, L2 and L3. The rest of the circuit is in simple cascade with CC. Therefore, the last connect code (CAX AA EE), connects C1 (AA), L1 (BB), CC, C2 (DD) and L4 (EE) in cascade. This completes the circuit which is now contained in two-port AA.

The OUTPUT statement AA(50) directs the program to analyze the circuit in AA with 50 ohm source and load terminations. And finally, the SWP statement indicates an analysis sweep in 10 MHz increments between 50 and 950 MHz.

Output from the program can be viewed in four different analysis windows via the pull-down View menu. Representative output for the broadband amplifier above is displayed in Figures 6, 7, 8 and 9. These are the Smith Chart, Plot, Results and Stability windows respectively.

The cluster of curves and frequency markers around the 50 ohm point at the center of the Smith Chart in Figure 6, indicates a good impedance match over the entire band. Figure 7 shows forward gain and phase versus frequency. Here it can be seen that the phase is quite linear with frequency and that the gain remains flat at 14 ±0.75 dB over the 900 MHz band. Figure 8 is a tabular numerical display of all 18 analysis indicators. From this

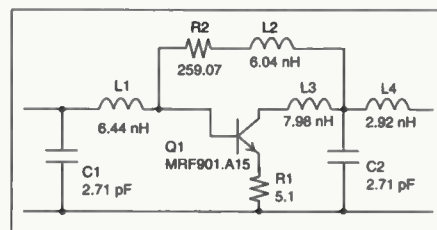
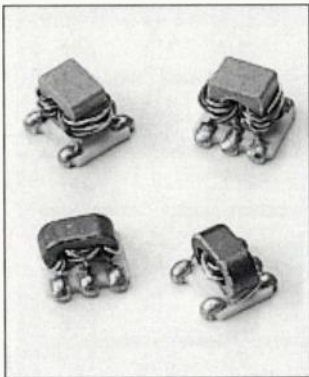


Figure 5. Broadband amplifier example.



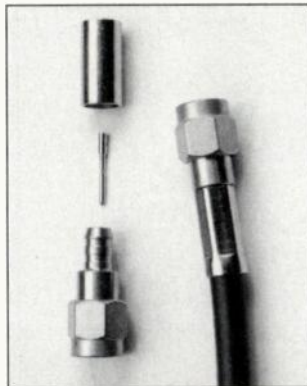
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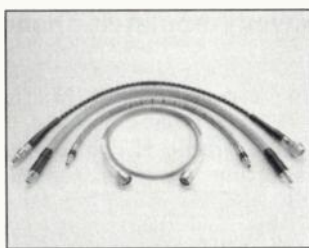


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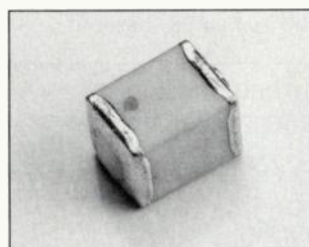


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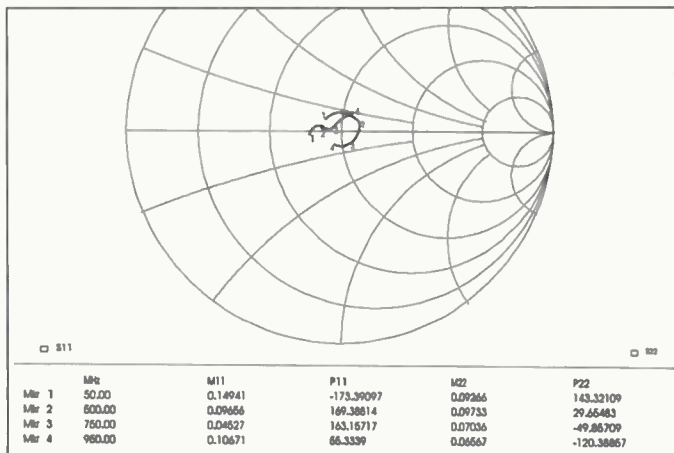


Figure 6. LINC output showing magnitude and phase of S11 and S22 at various frequencies.

data we can get numerical values for return loss (S_{11} and $S_{22} > 20$ dB), VSWR (1.2), input/output impedance (42/58 ohms) and gain (13.8 dB), etc. Finally, the stability analysis of Figure 9 shows that all stability circles completely enclose the Smith Chart and that the stable region is the entire input/output reflection coefficient plane. This ensures unconditional stability over the band.

One of the most important tools in this program is the ability to tune any circuit component while observing the effects it produces on any circuit response or performance parameter. Tuning is straight-forward and easy to use. All parts entered into the circuit file with a question mark preceding their value are marked for tuning. Any number of parts can be ganged together with the EQU (equate) statement so that tuning one will tune them all. Any part so marked can be selected for tuning by a pull-down Tune menu in the Plot Window. Once selected, a part can be tuned up or down in value by simply tapping the up or down arrow keys. The tuned response is shown as dashed lines while the initial response remains on the screen as a solid curve. This allows an instant visual comparison

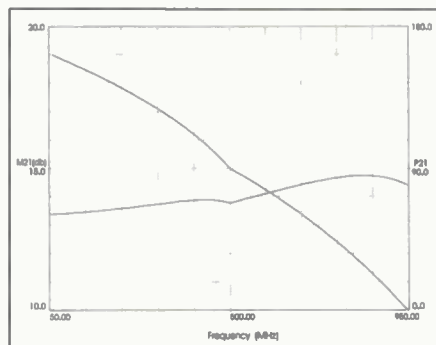


Figure 7. LINC output showing rectangular plot of magnitude and phase of forward gain.

between initial and tuned responses.

Significant insight into circuit behavior can be gained when the tuning simulation is as quick and interactive as it is in this program. The sensitivity of a circuit to component tolerances can be quickly evaluated by tuning a part's value and noting the degree of movement in the response. Tuning is not limited to adjustment of component values. Component parameters such as Q, transformer turns ratio, transmission line length and characteristic impedance etc., can also be tuned. The value of interactive circuit simulation becomes apparent when noting that these parameters would be virtually impossible to tune on the bench.

System Requirements

The following system requirements are not a minimum specification but only a suggestion for good windows performance. The program will run on a 4 megabyte 386SX machine with degraded performance.

486-DX processor (or Pentium™), local bus desirable (math coprocessor required for adequate performance), VGA monitor (standard 640x480), windows accelerator helpful, 8 Meg bytes RAM, Windows compatible graphics

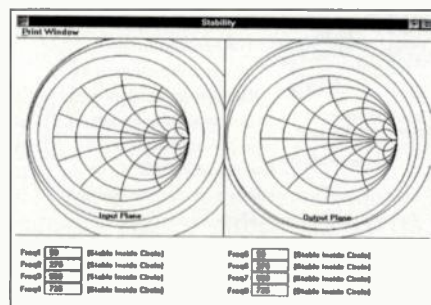


Figure 9. LINC output showing stability circles for the example broadband amplifier.

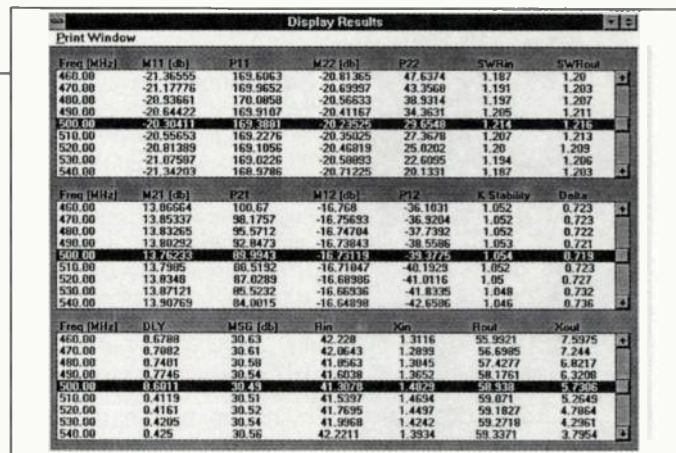


Figure 8. LINC output showing tabulated analysis results.

printer for hardcopy output, Microsoft Windows™ 3.1 or later.

LINC is available through Argus Direct Marketing. To order, see the ad on page 130. **RF**

References

1. Randall W. Rhea, *Oscillator Design and Computer Simulation*, Prentice Hall, Englewood Cliffs, NJ, 1990.
2. Tri T. Ha, *Solid-State Microwave Amplifier Design*, Krieger Publishing Company, Malabar, FL, 1991 (reprint). Chapter 1, section 1.5 "Interconnections of Networks". Chapter 2, "Scattering Matrix" and "Analysis of Two-Port Networks".
3. Max W. Medley, Jr., *Microwave and RF Circuits: Analysis, Synthesis and Design*, Artech House, Inc., Norwood, MA, 1993. Chapter 1: Two-port representation of circuit components, network connections and S parameters. Chapter 2: Construction of two-port networks from circuit elements, network analysis and characterization.
4. Guillermo Gonzalez, *Microwave Transistor Amplifiers: Analysis and Design*, Prentice-Hall, Englewood Cliffs, NJ, 1984. Chapter 1, "Representation of two-port networks". Chapter 3, section 3.3, on stability considerations and stability circles discusses the "K" and "Delta" stability indicators and what stability circles are.

About the Author



Dale Henkes is a Project Engineer in the Analog Signal Processing group at Philips Consumer Electronics Company in Knoxville, Tennessee. He is currently working on UHF receiver design for FCC part 15 applications. He received a BS degree in Engineering from Walla Walla College in 1977, and has recently completed graduate courses in EM fields and microwave networks at the University of Tennessee. He can be reached at 1916 Plumb Ridge Rd., Knoxville, TN 37932, or by telephone at (615) 521-3423.

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from the editors of **RFdesign**

RF Connectors see Changing Needs

This month's Product Forum highlights market changes affecting the RF connector industry.

RF Industries, Ltd.

Total connector shipments to the U.S. market will exceed \$8 billion by year-end 2000 according to a study by Leading Edge Report. Coaxial connectors shipments will amount to \$820 million or 10 percent of the overall market. The anticipated growth in these vertical markets is approximately \$100 million per year for the next five years.

The driving force behind this tremendous growth is the increased demand for communications equipment, computers and their networks, and general microwave equipment. In many cases the connector used will not be standard. It will, however, be RF and connector manufacturers will strive to provide new offerings. Both price and delivery will become extremely competitive in coming years.

Penstock

In the ever-changing world of coaxial RF connectors, smaller size and higher frequency requirements are driving the industry. The connector companies that comply to customer demands are growing at rates of up to 30 percent, along with the increasing "wireless" market. The \$350 million (per calendar year 1994) RF connector market is changing so rapidly because customer design to market cycles have gone from 2-4 years to 6 months-1 year. Most connector companies are used to dealing with the military/industrial marketplace and find the consumer marketplace extremely fickle.

Distribution is increasingly being used by OEMs to meet new commercial design-in requirements because of the short design cycles. Distributors are participating in high volume business that would have been OEM directed only a few years ago.

AMP Incorporated

We expect to see good growth opportunity in the RF market over the next few years with major focus in the telecommunications market and the key applications of the cellular telephone and the cellular base station markets. PCN/PCS and the wireless communications markets are also

expected to be major contributors to the growth. A similar growth is occurring in the medical industry for applications in diagnostic and imaging equipment.

Over the past few years, the RF coax market has experienced slow growth in both the European and Pacific Rim regions. However, during the past year, we have seen improvement in RF activity as the economies of these regions are improving.

Today's technology is generating packaging requirements for smaller and smaller components. These trends toward miniaturization and surface-mount technology are playing very important roles in the RF market.

Andrew Corporation

DIN connectors for coaxial cable, once only popular in Europe, are now gaining in popularity in the U.S. In many cellular, LMR, trunking and PCS/PCN applications, DIN connectors are replacing the time-tested N connector.

Intermodulation (IM) performance of passive devices is more critical than ever. DIN connectors, when used in conjunction with coaxial cable constructed with solid inner and outer conductors, provide superior IM performance. They do not use ferromagnetic materials, such as nickel, which can produce IM. In addition, higher contact pressures, resulting from their robust design, break through thin surface oxide layers, resulting in a reduction of IM generation. DIN connectors are now available in sizes ranging from 1/4" superflexible cable to 2 1/4" low density foam cable.

E.F. Johnson

By prohibiting the use of standard coaxial connectors for antenna mounting, the FCC has greatly affected the growing wireless market. E.F. Johnson Company recommends the standardization of reverse (left-hand) threaded SMA connector bodies for antennas to eliminate user error by mismatching the antenna. Unlike reverse thread connectors, reverse polarity contacts are not foolproof as they can be forced — causing damage to the connector and violation of FCC Rule 15. With snap-on connectors, an extended contact and insulator design is recommended to preclude forced or accidental connection with the center contact of a standard connector. The

cost-effectiveness of these recommended options greatly outweigh the total redesign costs of reverse polarity contact mating options. You may contact E.F. Johnson Company for a detailed evaluation at 1-800-247-8256.

M/A-COM, Inc.

The commercial communications marketplace has been fueled by the global growth of wireless communications, and connector manufacturers have responded with smaller, more robust interconnects capable of handling a variety of miniature cables and frequency requirements.

Traditional suppliers of high reliability, rugged, mil spec type connectors have turned their technology expertise to smaller, higher frequency, robust designs with desired features like surface mount capability, pick and place assembly, ease of handling and test, and a very low cost. A range of products exist today that enable interconnections to take place on boards with mated height profiles ranging from 3.7 mm to 9.5 mm handling frequencies from 2 to 8 GHz, accommodating a variety of miniature cables.

Coaxial Components Corp.

Commercial application in all facets of wireless communications is driving increases in the miniature and micro miniature RF connector markets. The key to success for the RF connector manufacturer in the commercial market will be to reduce package size, and cost while increasing volume. As new requirements for the commercial market continue to emerge we have some reservations as to how long market growth will continue for some items; we feel it is inevitable that functions of some components using the RF connector will become hybridized and eliminate the use for those RF connectors in the future. Coaxial Components Corp. manufactures a complete line of RF connectors ranging from DC to 26.5 GHz.

The Phoenix Company of Chicago Inc.

The market for interconnect is growing but the amount of interconnect per unit produced is shrinking. RF and microwave separable interconnect development is being driven by high volume commercial applications such

as cellular, personal communications, and GPS. Traditional manufacturing technologies like screw machine and stamping have been enhanced in order to meet the requirements for high performing, downsized components.

The Phoenix Company of Chicago has created its PMMX series surface mount RF connection system to meet these new demands. This system will accommodate standard, low loss, flexible RG and semi-rigid cable designs. The surface mount cube is machined from brass in order to maximize RF performance. This connector series allows for 500 mating cycles and is rated to 6 GHz.

As system designers are striving to eliminate separable interconnect where possible, the connector maker is challenged to produce a small size, high performance, and low applied cost connector.

Lucas Weinschel

In our opinion, the Blind-Mate RF connector market is growing but at a modest pace. Its key applications are in automatic test equipment and systems, repetitive testing of RF modules and those areas of aerospace and telecommunications where package densities are high and conventional connections via connector coupling nuts are inaccessible.

We see this product category to be very price sensitive since major business developments are rapidly shifting to the aggressively competitive commercial arena. High reliability, environmentalized versions of blind-mate connectors will still have their place in military applications, where performance and ease of use will be the dri-

ving factors - not price. A particular area of growth will be in the surface mounted blind-mate connectors with higher operating frequencies than presently available.

Huber + Suhner, Inc.

Coaxial RF connectors are playing a more significant role in the electronics marketplace due to more stringent electrical and mechanical requirements, lower unit and installed cost, quality levels measured in parts per million, and service levels equivalent to L.L. Bean. Critical customer needs

today include microminiaturization, higher performance, and vendor and cost reduction initiatives.

Huber + Suhner supports all RF connector market segments worldwide, and is positioned to support the exploding wireless communications segment from the cell site to the PDA. As an antenna, cable, component, and connector manufacturer, Huber + Suhner understands the total system performance requirements to ensure optimum performance. Due to this, Huber + Suhner has outpaced the market growth rate over the last five years. **RF**



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RF literature

Crystal Oscillators

Milliren Technologies' short form catalog contains specifications for over thirty of MTI's low noise, high precision quartz oscillators, including OCXOs, TCXOs and VCXOs. The catalog is available from the factory free of charge.

MTI - Milliren Technologies, Inc.
INFO/CARD #216

SAW Products and Applications

Sawtek's 1995 product catalog includes product listings, a section of specifying a custom SAW filter, application-specific SAW devices, and PC board layout tips to assist the RF designer. The 60-page catalog also has a section that explains the fundamentals of SAW transversal filters for those who are unfamiliar with SAW technology.

Sawtek Incorporated
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Ceramic Chip Capacitors

The 20-page catalog from Johanson Technology Inc. describes in detail a wide variety of RF/Microwave ceramic chip capacitors. The catalog features the Johanson LASERtrim™ surface mount tuning capacitor, a laser adjustable monolithic ceramic surface mount devices for precise tuning of RF circuits. Also included are a variety of high-Q types, single layer microwave chip capacitors as well as a section on high voltage capacitor arrays.

Johanson Dielectrics, Inc.
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Power Amplifiers

A 6-page short form catalog entitled RF Power Amplifiers is now available from ENI. The catalog contains primary specifications for thirty different broadband and

pulse power amplifiers, with power outputs ranging from 3 to 8000 Watts and frequency coverage from 9 kHz to 1000 MHz. ENI amplifiers feature RF stability, +13 dBm overdrive protection and solid state design.

ENI
INFO/CARD #213

MMIC Amplifier Data Sheet

FEI has released a two-sided data sheet on its Model AH-B102D-3 series of low cost, cascaded HBT GaAs MMIC amplifiers. Drawings illustrate size, packaging, a recommended bonding scheme, typical biasing configurations and a simplified schematic of the MMIC.

FEI Communications, Inc.
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Microwave Components

RLC Electronics, Inc. has available its 1995 full-line catalog. This edition features surface mount packages for switches and filters, waveguide bandpass filters, and an expanded line of power dividers.

RLC Electronics, Inc.
INFO/CARD #211

Data Book

RF Monolithics, Inc. has published a comprehensive data book detailing its 210 standard products. The 528-page data book, available at no cost, details the company's products in three major product categories: low-power wireless, high-frequency timing, and telecommunications. The book features short-form catalogs, individual product specifications and case dimensions.

RF Monolithics, Inc.
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Wireless Products

K & L Microwave's latest catalog fea-

tures its wireless product line. The Wireless Products catalog highlights K & L's capabilities and products. Products covered include: ultra high "Q" base station filters and duplexers, wireless transmit and receive filters, mobile duplexers, Dielectric resonator and ceramic filters, the KEL-film™ disc, integrated subassemblies, and coaxial switches.

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Measurement Instruments

Precision PC instruments are included in Guide Technology's measurement instruments catalog. Some instruments included are a 2 GHz time interval analyzer, a 200 MHz digital oscilloscope, a 5 1/2-digit multimeter, and a 1.3 GHz counter.

Guide Technology Inc.
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Communications Equipment

Andrew Corporation has published a new edition of its general catalog. The 620-page, full-color catalog has major sections on microwave systems; broadcast and satellite earth station antenna systems; HELIAX® coaxial cable, connectors and accessories; waveguide; towers and equipment shelters.

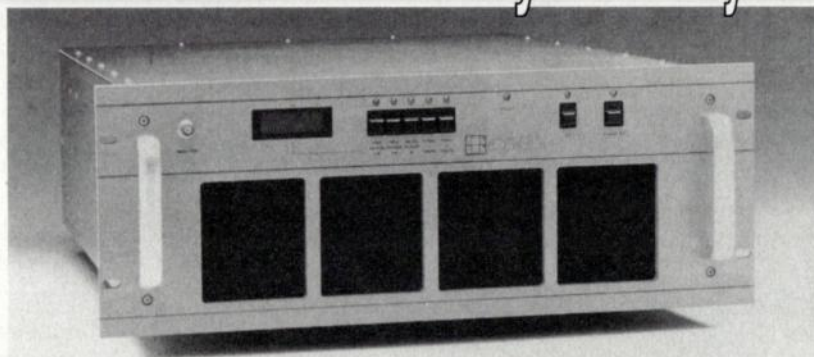
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EMI Filter/Chip Inductors

Catalog no. E-06-B has been released from Murata Electronics North America. The 110-page catalog contains detailed information on Murata's line of surface mount EMI filters, surface mount inductors, EMI leaded filters, EMI filter connectors, AC EMI filters, noise filters, and EMI filter/inductor design kits.

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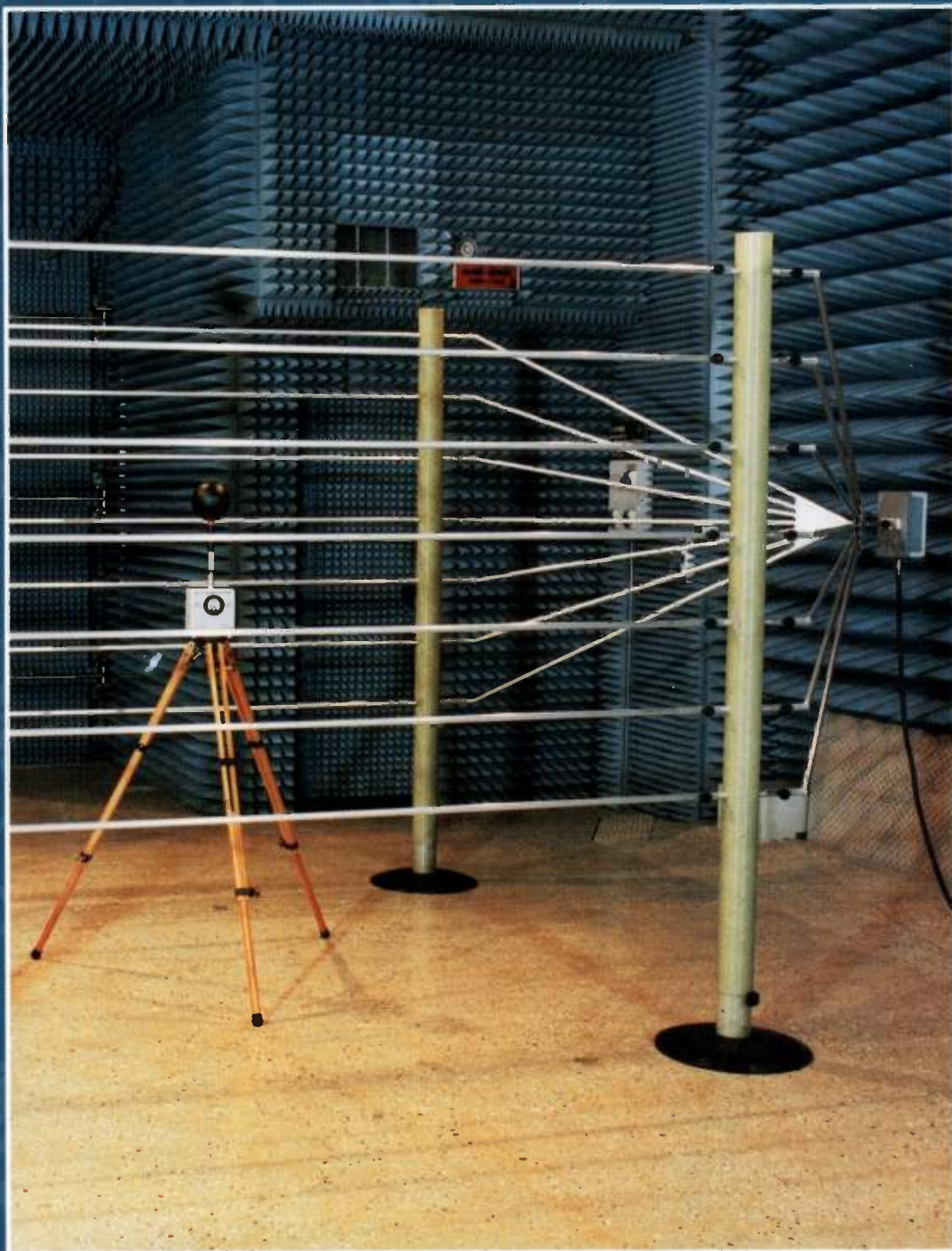
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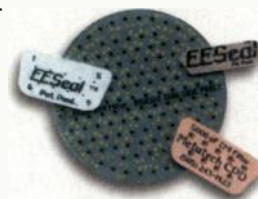
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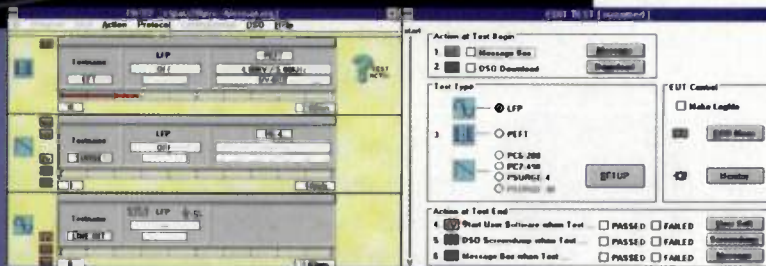


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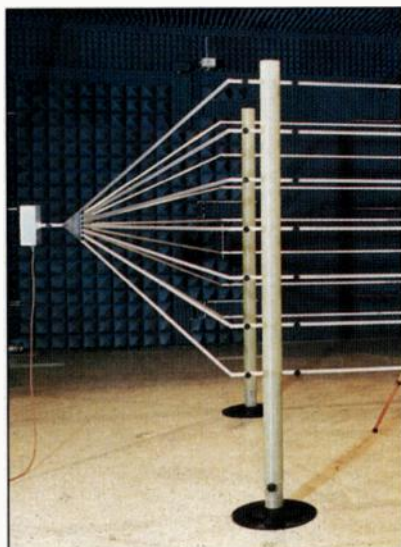


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INFO/CARD 50

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The photo on the cover of this issue of *EMC Test & Design* shows the HMC antenna described in the article beginning on page 20. This antenna was developed for radiated immunity testing to MIL-STD-461D, tests RS101 and RS103, as well as to IEC 801-3.

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EMC

TEST & DESIGN

FEATURES

Choke Design for DC Motor Noise Suppression

14

Stuart Gordon and Patrick Cattermole

Meeting regulatory standards for conducted and radiated emissions is required for most electronic equipment. When DC motors are used, the arcing between the brushes and commutator creates broadband noise that must be filtered. This article discusses the ferrite and iron powder material options for common mode and differential mode chokes, with guidelines on the proper configuration for each choke type, and for low and high frequency ranges.

A New Tool for Immunity Measurements up to 200 MHz

20

Lorenzo Carbonini

This article describes the theoretical and experimental analysis of a multiconductor transmission line for radiated immunity measurements. The Horizontal polarization MultiConductor (HMC) antenna generates both electromagnetic fields (10 kHz - 200 MHz) and magnetic fields (up to 1 MHz). The antenna generates fields in a large test region of 1.8 m × 1.8 m × 3 m.

De-Certification of PCs and Peripherals Proposed by FCC

35

Terry G. Mahn

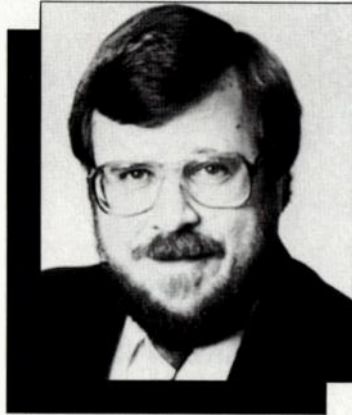
The FCC has proposed streamlined authorization procedures for personal computers and PC components that are intended for the home market. ET Docket 95-19 outlines the Declaration of Conformity (DoC) that manufacturers would use in place of the current certification process. The scope of the proposal and details of implementation are covered in this regulatory update.

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Editorial

It's Time to Wake Up to Design-for-Compliance



Why do I still hear the horror stories? You know the kind — products that arrive at the test engineer's office that are supposed to start shipping in three weeks, but no one knows whether they will pass FCC Part 15. Or, despite extensive consultation on the early design, so many changes were made that all the EMC-robust techniques have been corrupted by the time the product leaves the development lab. Or, after successful EMC testing, the product manager wants to make some last-minute changes. Aaaaargh!

I heard the best story just a few days ago: A product was delivered to the EMC test engineer with a short timetable before shipping — with the casual remark from the project engineer, "We know there's a radiation problem in the clock circuit. In fact, we've known about it for months, but we've been too busy getting the firmware debugged to look into it."

Despite many good efforts to avoid these kinds of problems, they still happen with alarming regularity. When will company management get the message that EMC is part of the whole engineering mission, not an add-on at the end of the line? I know they understand how irate stockholders can get when products don't make it to market on time; it's time someone spelled out the strong connection between "doing the job right" and making a profit!

In today's commercial/consumer electronics environment, time-to-market and cost are the most important factors in product development. These requirements can only be met with a commit-

ment to doing all the right things — in product definition, functional design, manufacturability, reliability, *and EMC compliance*. Why is it so obvious to so many of us, and impossible to grasp for so many others?

I marvel at some of the small companies that create top-notch products on a shoestring budget and small staff. They have a natural environment that encourages (requires!) teamwork and ongoing communication among all job functions. After all, the design engineer might also do manufacturing, and the V.P. of engineering might be directly responsible for EMC compliance. Each employee might be a stockholder, too.

This entrepreneurial atmosphere disappears in many larger companies, so other methods are substituted, with mixed success. In some big companies, design teams have become extremely effective in creating new products on time and within budget. Other companies have become the subject of the cartoon strip Dilbert — their ineffective plans ruthlessly parodied to the delight of anyone who has worked in an inflexible bureaucracy.

This isn't the first time I have exhorted industry to wake up to need for EMC as an integral part of engineering development. And, I will keep preaching on this subject from my editorial soapbox as long as necessary! It's too important to ignore.

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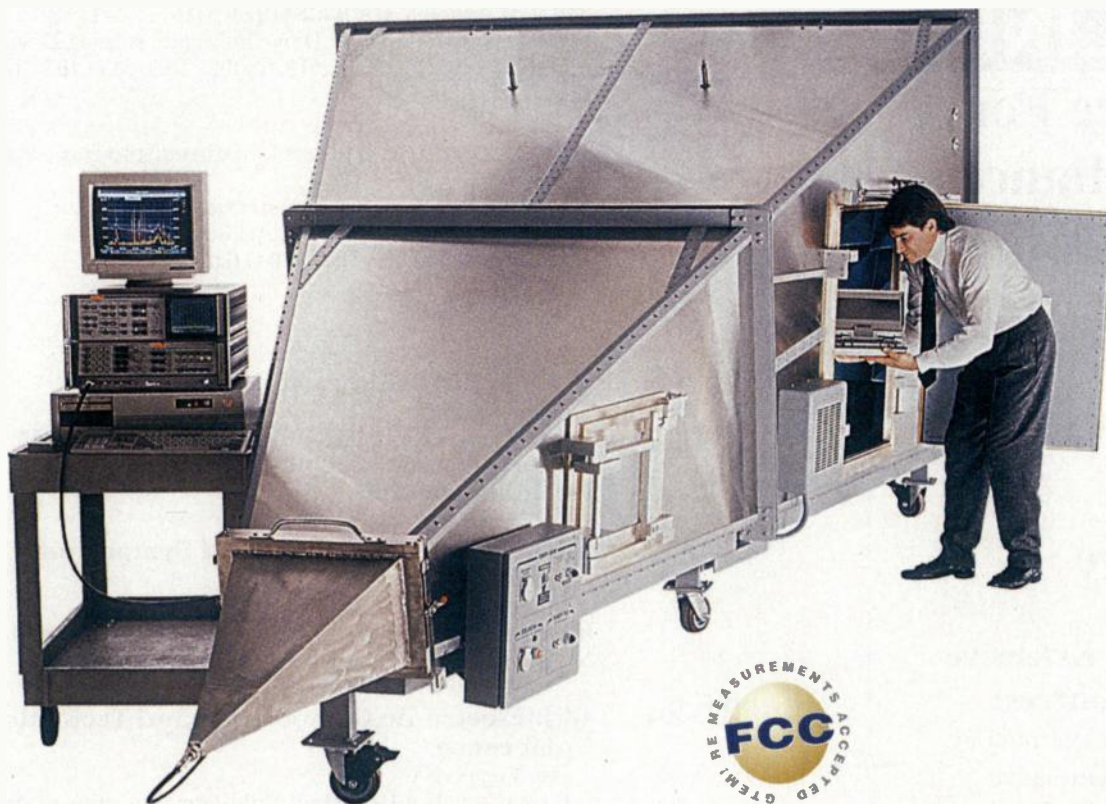
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■ Calendar

April

- 23-26 IEEE Instrumentation/Measurement Technology Conference**
Waltham, MA
Information: Robert Myers, 3685 Motor Ave., Suite 240, Los Angeles, CA 90034. Tel: (310) 287-1463. Fax: (310) 287-1851, or Dan Sheingold, Analog Devices, P.O. Box 280, Norwood, MA 02062. Tel: (617) 461-3294. Fax: (617) 329-1241.
- 25-27 International Wireless Communications Expo 95**
Las Vegas, NV
Information: IWCE 95, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828-0420. Fax: (404) 618-0441.

May

- 15-19 IEEE/MTT-S Microwave Symposium**
Orlando, FL
Information: 1995 IEEE Symposium, c/o Horizon House Publications, 685 Canton Street, Norwood, MA 02062. Fax: (617) 762-9230.
- 19 Fifth-Annual IEEE Regional Symposium on EMC**
Boulder, CO
Information: Bob German, Henry Ott Consultants, 1410 Moss Rock Place, Boulder, CO 80304. Tel: (303) 444-2472.
- 21-24 45th Electronic Components and Technology Conference**
Las Vegas, NV
Information: Jim Bruorton, Publicity Chairman, 1995 Electronic Components and Technology Conference, c/o KEMET Electronics Corporation, P.O. Box 5928, Greenville, SC 29606. Tel: (803) 963-6621. Fax: (803) 963-6521.
- 31-2 1995 Virginia Tech Symposium on Wireless Personal Communications**
Blacksburg, VA
Information: Jenny Frank, Administrator, Mobile and Portable Radio Research Group. Tel: (703) 231-2958.

June

- 1-2 CEM 95: The 3rd Portuguese Seminar on Electromagnetic Compatibility**
Lisbon, Portugal
Information: Silicon Electronica E Telematica, Edificio Pascoal de Melo, Rua Pascoal de Melo, N. 3, 1100 Lisboa, Portugal. Tel: 8151234. Fax: 8130796.
- 6-8 Medical Design and Manufacturing Trade Show**
New York, NY
Information: Canon Communications, Inc., 3340 Ocean Park Blvd, Suite 1000, Santa Monica, CA 90405-3216. Tel: (310) 392-5509. Fax: (310) 392-4920.

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United Kingdom, 71-932-1505 (Fax 71-931-0175)

15/FI Wilson House, 19-27 Wyndham Street Central, Hong Kong, 521-2333
(Fax 810-1965)

Royal Trust Tower, Toronto-Dominion Centre, Toronto, Ontario M5K 1N4,
Canada, 416-361-7778 (Fax 416-361-7659)

■ Courses

Standards and Calibration Laboratories: Principles and Practices

May 8-12, 1995, Washington, DC

Electromagnetic Interference and Control in Modern Communications Systems

May 15-19, 1995, Washington, DC

Information: The George Washington University,
Continuing Engineering Education, Academic Center,
Room T-308, 801 22nd Street, N.W., Washington, DC 20052.
Tel: (202) 994-6106 or (800) 424-9773. Fax: (202) 872-0645.

Grounding and Shielding Electronic Systems - How to Diagnose and Solve Electrical Noise Problems

April 18-19, 1995, Boston, MA

Circuit Board Layout to Reduce Noise Emission and Susceptibility

April 20, 1995, Boston, MA

Information: Continuing Education, University of Missouri-
Rolla, Rolla, MO 65401-0249. Tel: (314) 341-4132 or
(314) 341-4200. Fax: (314) 341-4992.

Electronic Design Techniques and Analysis Required to Meet Electromagnetic Compatibility Requirements

May 3-4, 1995, Novi, MI

Advanced EMC Printed Circuit Board Design

May 5, 1995, Novi, MI

Information: JASTECH, James P. Muccioli, P.O. Box 3332,
Farmington Hills, MI 48331. Tel: (810) 553-4734.

Electromagnetic Compatibility Engineering: EMC Design and EMI Mitigation

May 22-23, 1995, East Brunswick, NJ

International EMC Standards, Requirements, Measurements, and the European Union Approach

May 24-26, 1995, East Brunswick, NJ

Information: Registrar, The Center for Professional
Advancement, P.O. Box 1052, East Brunswick, NJ
08816. Tel: (908) 613-4500. Fax: (908) 238-9113.

How to Meet the Immunity Requirements of the European EMC Directive, and Get Your CE Mark

May 3, 1995, New York, NY

May 5, 1995, Boston, MA

May 11, 1995, Rochester, NY

Information: The Boxleitner Group, 126A Pleasant Valley
Street, #414, Methuen, MA 01844. Tel: (508) 687-4486.

Producing an EMC Technical Construction File

April 25, 1995, Luton, UK

May 3, 1995, Warwick, UK

May 4, 1995, Reigate, UK

Automotive EMC Approvals

April 26, 1995, Warwick, UK

Information: Nigel Harvey, SGS EMC Services, Hutton
Building, St. Michael's Way, Sunderland SR1 3SD, UK.
Tel: 0191-515-2663. Fax: 0191-515-2670.

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■ News

ICC Accredited by NVLAP and NEMKO

International Compliance Corporation (ICC) has received accreditation in electromagnetic compatibility and communications under the National Voluntary Accreditation Program (NVLAP). ICC's scope of accreditation includes conducted and radiated emissions in accordance with FCC CFR 47 Part 15 Subpart B - Digital Devices.

ICC also received an EMC Laboratory Authorization from NEMKO. NEMKO is a Nordic independent testing and certification body offering documentation on electrical safety/radio interference in relation to ITE/office equipment.

CKC Expands

CKC has completed expansion which added fully-anechoic chambers to all regional operations. They now have a 10'x20'x28' chamber in Hillsboro, OR and a 10'x10'x20' chamber in Brea, CA providing testing services. The new, fully-anechoic chambers utilize ferrite-tile, RF absorption technology. They comply with the new European chamber calibration standards for radiated immunity testing between 27 MHz and 1 GHz. Both of the facilities are testing to the latest immunity standards as called out by the Generic Standards EN50081-2 and EN50082-2 and the emissions standards as called out by

the Generic Standards EN50081-1 and EN50082-1.

ESD Control Certification Program

The ESD Association has announced the beginning of the ESD Control Certification Program. In order to become qualified as an Electrostatic Discharge Control (ESDC) Engineer, the applicant must meet the following requirements: nine years of experience as an engineer in the area of ESD control, (Education can be substituted for some experience.); endorsements by three engineering professionals in the area of ESD control; submission of ten test questions for use in future ESDC engineering qualification tests; submission of a summary of experience; satisfactory passing of the ESDC Engineer's qualification test administered by NARTE; and membership in the ESD Association. To become an ESDC Technician the applicant must meet these requirements: six years of experience as an engineer in the area of ESD control (Education can be substituted for some experience.); endorsements by three engineering/technical professionals in the area of ESD control; submission of ten test questions for use in future ESDC engineering qualification tests; submission of a summary of experience; satisfactory passing of the ESDC Engineer's qualification test administered

by NARTE; and membership in the ESD Association.

Wyle Laboratories Changes Name

Wyle Laboratories has change its name to Wyle Electronics. This change will more accurately reflect the company's business activities following the sale of its Scientific Services and Systems Group.

ETC Purchased by MPB Technologies

The Alberta Research Council (ARC) has signed an agreement with MPB Technologies Inc. for the transfer of the Electronics Test Center (ETC). MPB Technologies resumed full responsibility for ETC operations in Alberta on January 3, 1995. ETC's new owner will honor existing contractual commitments and maintain current accreditations. The agreement also calls for the test facility to be moved to Calgary. The ETC is an extension of MPB Technologies' EMI/EMC test measurement facility in Ottawa. The two facilities, along with its Montreal-based activities, allow MPB to provide complete measurement services nationwide.

Aerovox Establishes Electrolytic Technical Center

Aerovox Incorporated has established an electrolytic technical center at its Aero M Group plant in Huntsville, AL. The test lab for aluminum electrolytic capacitors is dedicated to diagnostics, testing and material product development. The facility brings together the technical and production expertise of two Aerovox aluminum electrolytic manufacturing divisions, Aero M and BHC Aerovox Ltd. of England.

PowerCET Expands

PowerCET Corporation has expanded its base of operations to both coasts with the opening of a new office in Gainesville, FL. The new Florida office offers expanded services to PowerCET's clients for education and training, on-site problem investigations, equipment and systems evaluations, and lightning and surge protection systems consulting services.

TUV Rheinland of North America Opens EMC Testing Facility

TUV Rheinland of North America has opened a new EMC testing facility on the grounds of the company's North American headquarters in Newtown, CT. Products that can be tested for electromagnetic compatibility at this facility include home appliances, computers, peripherals and RF remote controls, among many others. TUV Rheinland can conduct the appropriate testing and help compile the necessary Technical File in order to obtain the CE Marking. The new facility is fully operational and allows for complete compliance testing for domestic (FCC) emissions requirements and those of the EMC Directive 89/336/EEC.

The new EMC building is approxi-

mately 2500 square feet including a semi-anechoic chamber. The chamber has a six-foot turntable with a 2000-pound capacity. The Open Area Test Site is a fully heated and air conditioned, vinyl fabric air-supported structure with a 10-foot diameter, 2000-pound capacity turntable.



Choke Design for DC Motor Noise Suppression

By Stuart Gordon and Patrick Cattermole
MMG North America

Stuart Gordon is Vice President of Engineering and has 30 years experience at MMG North America. He holds a B.S. in Ceramic Engineering and a Doctor of Science, Ph.D. Patrick Cattermole is Senior Applications Engineer. He has been with MMG for 10 years and has a BSEE degree.

The authors may be reached at MMG North America, 126 Pennsylvania Avenue, Paterson, NJ 07503-2512; tel: (201) 345-8900.

Due to governmental regulations, there is an increasing need to limit noise emissions from DC motors, as well as from other electrical and electronic devices. Ferrite and powdered iron can be very effective as core materials for differential and common mode chokes for use in controlling both conducted and radiated interference. Through selection of material type, core geometry, and winding configuration, these chokes can be designed to achieve optimum noise suppression performance.

Introduction

DC motors generate a significant amount of radio frequency interference (RFI), due to arcing at the brush-to-commutator contact points. An electrical spark produces a wide spectrum of electromagnetic emissions. There is a need to prevent these emissions from being conducted or radiated from the motor, due to their deleterious affects upon high speed digital and general communications equipment. The elimination of these spurious signals is a requirement in most industrial markets including Europe, Japan and the United States.

Conducted emissions can be controlled by a combination of capacitors and differential and/or common mode chokes. Radiated emissions are usually suppressed with the aid of differential mode chokes. Some of the factors to be considered in the design of both common mode and differential mode chokes for noise suppression in DC motor applications will be examined below.

Noise Emissions From DC Motors

Electric currents are initiated and abruptly terminated at the brush-commutator interface of a DC motor. This occurs both on a macroscopic and microscopic level. On the macro level, a current initiates as the brush makes contact with a particular commutator pad and terminates when contact with the pad ceases. On the micro level, current paths are initiated and terminated as the brush wipes across the surface of the pad. A rapidly changing current generates high frequency components, which in the case of a typical DC motor can produce an emission band spanning a frequency range from 10 kHz to over 100 MHz. In most applications, this noise must be suppressed as its presence may cause interference to electronic data transmission.

The potential hazard to the operation of sophisticated electronic equipment due to electromagnetic noise emanating from electronic and electrical devices, including DC motors, has ne-

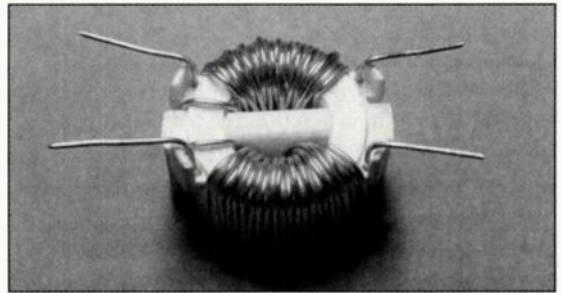


Figure 1. Toroidal common mode choke.

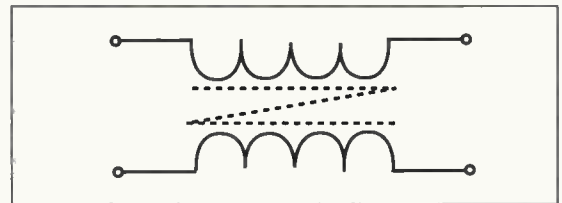


Figure 2. Schematic of a common mode choke.

cessitated the development of a body of regulations governing their electromagnetic interference characteristics. Measures must be taken to insure that a device incorporating a DC motor complies with applicable regulations. A computer peripheral containing motor drives, such as a printer or data storage drive, is an example of a device falling under such government regulation. Emissions from a computer system and its peripherals are regulated by the FCC in the United States and by similar regulatory agencies in other countries such as VDE in Germany and CISPR in the European Community. In general, the various regulations all define frequency bands and acceptable levels for both conducted emissions, that is, those carried along electrical wiring connected to the device, and radiated emissions, which travel through free space from the device. In the case of the FCC in the USA, conducted emissions are regulated in the frequency band from 150 kHz to 30 MHz. Radiated emissions must be controlled at frequencies above 30 MHz and reaching into the GHz region.

Suppressing Conducted and Radiated Emissions

Conducted noise emissions are propagated over the electrical power connections to a DC motor. Inductors and capacitors can be used in combination, to shunt and impede these undesirable currents. Noise currents are quite often common mode, that is, they are in phase on both power conductors, but differential mode interfer-

■ Choke Design

ence may also be conducted along the power line from the motor. Both of these types of conducted noise produced by a DC motor may require suppression. If only common mode noise is present, a common mode choke will suffice. However, if both common and differential mode noise is being conducted, a differential mode choke (a larger and more costly component) must be employed.

Radiated noise emissions are RF signals emanating from an electrical device, such as a DC motor, and traveling through free space. Such signals may possibly interfere with the proper operation of telecommunications equipment, radio and television equipment (both commercial and military), medical and scientific apparatus, information technology equipment, as well as other types of electronic devices. Generally, governmental regulations cover a frequency band starting at 30 MHz and extending up to 300 MHz. In some cases, the regulated band may extend to higher frequency, possibly into the GHz region. Differential mode chokes are the most effective means of suppressing radiated emis-

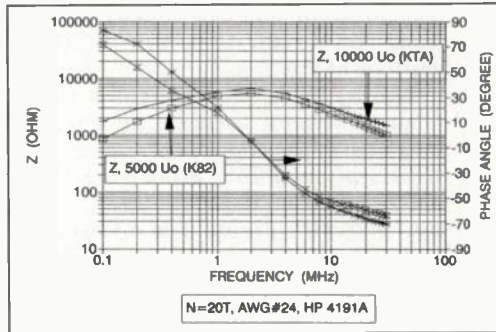


Figure 3. Impedance and phase angle versus frequency for the common mode choke.

sions. Such a choke should be placed as close to the motor as possible and should be designed such that its impedance is maximized over the generated noise frequency band.

Design Considerations For A Common Mode Choke

In the case of in-phase conducted noise currents, a common mode choke is an effective way to provide electromagnetic interference (EMI) protection. The frequency band for conducted emissions starts in the kHz region,

and extends up to 30 MHz. At the low end of the spectrum, relatively large choke inductance values are necessary in order to obtain significant AC impedance values. Manganese-zinc soft ferrite is most suitable for use as a core material for common mode chokes because it exhibits relatively high magnetic permeability at low frequencies. This results in high impedance at the low end of the frequency range, as well as efficient coupling. In the case of a single winding, a typical high permeability core would readily saturate

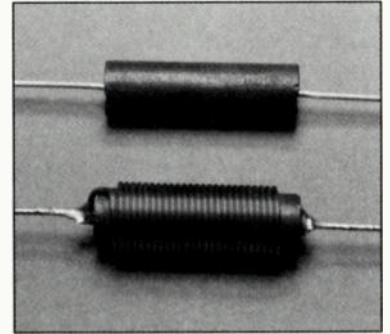
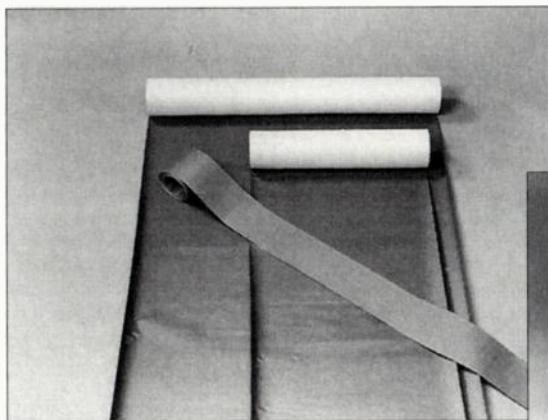


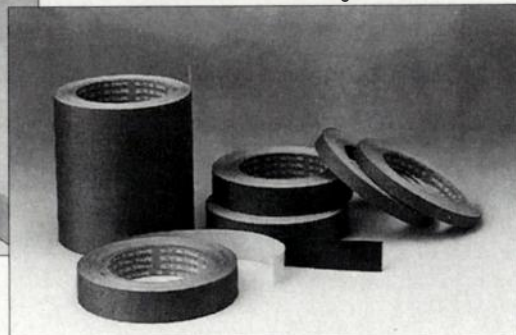
Figure 4. Photo of a differential mode choke.

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Choke Design

with a relatively low DC current running through it. However, by winding the core utilizing two sets of windings of equal turns but having opposite directions, the fields generated by the DC supply currents in each side of the line will cancel each other, and the core sees no net magnetic field. This allows a high impedance to be presented to the common mode noise currents without risk of saturation.

A toroid, or "ring" core is generally

used for common mode choke construction. The toroid shape takes full advantage of the high permeability of the ferrite material because it is a closed magnetic circuit without air gaps. In most cases, manganese-zinc ferrite with a permeability of at least 5000, but more often 10,000, is used as the core material. Figure 1 is an example of a typical common mode choke wound on a ferrite core, and Figure 2 is a schematic representation of the

device. Note that the two windings are physically separated by an insulating barrier, because of the significant DC potential between them.

Figure 3 is a graph of impedance and phase angle versus frequency for typical common mode chokes based on 5,000 (MMG K82 material) and 10,000 (MMG KTA material) permeability toroidal cores. Note that the impedance of the 10,000 permeability choke is significantly higher over the entire measured range, which may be of importance for some regulations.

As can be seen from Figure 3, the 10,000 permeability choke has an impedance of 1000 ohms or greater from 100 kHz to 30 MHz. An important aspect of the impedance versus frequency characteristic of this choke (and chokes in general) is self-resonance. Above the self resonance frequency, the device is capacitive, not inductive. Note in Figure 3 that the phase angle plot passes through zero and becomes negative at about 1.8 MHz. The choke is still effective beyond the self resonance frequency due to capacitive reactance in this region, but the impedance is decreasing, typical of the behavior displayed by a capacitor.

Design Considerations for a Differential Mode Choke

Differential mode chokes are the most suitable means for controlling radiated emissions in the 30 to 150 MHz

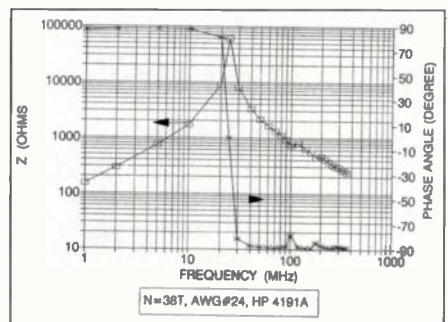


Figure 5. Impedance and phase angle versus frequency for a differential mode choke.

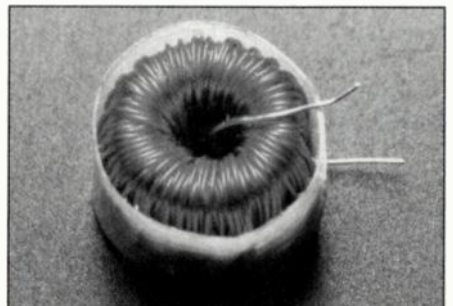


Figure 6. Powdered iron differential mode choke.

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Choke Design

frequency range. Unlike a common mode choke, where the windings are so arranged that the DC supply current to the motor produces no net magnetic flux in the core material, the differential mode choke must be able to withstand the magnetic field generated by the DC supply current flowing through it. This usually necessitates a magnetic component with a low effective permeability.

Since the flux density, B (in gauss), is equivalent to the product of the circuit permeability μ_c , and the magnetizing field strength, H in oersted, B is given by:

$$B = \mu_c H \quad (1)$$

The magnetizing field, H , may be expressed as:

$$H = (0.4\pi NI) / l_e \quad (2)$$

where N is the number of turns of the choke winding, I is the DC current and l_e is the effective path length of the core in centimeters.

The value of μ_c must be kept low enough to limit the peak flux density from exceeding the core material's saturation magnetization, for a given DC motor current, I .

A particularly efficient and cost effective method of fabricating a choke with low permeability is to use a ferrite rod as the core of a solenoidal winding. This magnetic circuit has a very large air gap, with lines of flux emanating from one end of the rod and returning to the other end through free space. The large air gap results in reducing the permeability of the magnetic circuit to some effective value, thus enabling the unit to handle relatively high DC fields. The effective permeability of such a choke, μ_c , may be defined as the ratio of its inductance L_e , to the inductance of the solenoidal winding without the core, L_a , and is given by:

$$\mu_c = L_e / L_a \quad (3)$$

The inductance of the choke in henries may then be approximated by:

$$L_e = (4\pi\mu_c N^2 A 10^{-9}) / l \quad (4)$$

where A is the cross sectional area of the coil, and l is its length in cm.

Equation 4 is accurate only in the case of a very long coil, but can be used as a first approximation for design purposes in many cases. Figure 4 is a photograph of a typical differential mode choke wound on a ferrite coil

form. The form is composed of a ferrite slug with axial leads bonded to the ends for winding termination.

As is the case with the common mode configuration, the choke resonates and becomes capacitive at a frequency which is dependent on its inductance and self capacitance. The impedance decreases above this self resonant frequency (SRF), but the choke still displays significant impedance well beyond the SRF.

The primary consideration when designing a differential mode choke intended to suppress radiated noise, is the position of the SRF. Impedance is greatest at this point, and the SRF should be made to coincide with the center of the maximum noise frequency band. For a given choke, self resonance occurs at a frequency at which the inductive reactance equals the capacitive reactance.

Capacitive reactance, X_c in ohms, is

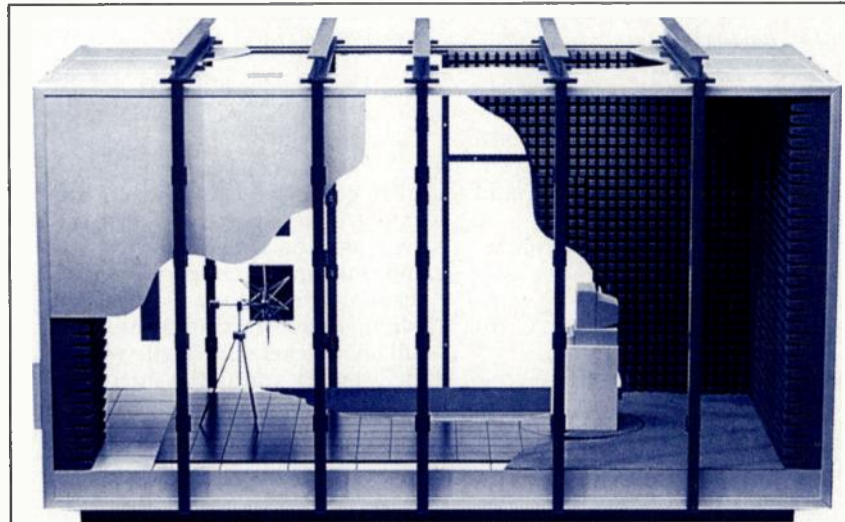


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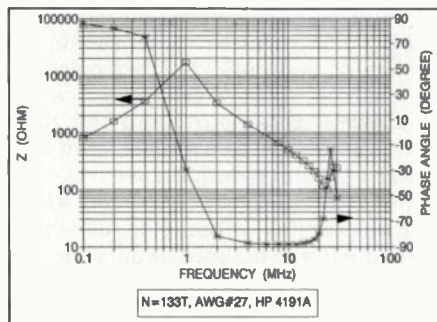


Figure 7. Impedance and phase angle versus frequency for a powdered iron differential mode choke.

given by:

$$X_c = 1/(2\pi fC) \quad (5)$$

where f is the frequency in MHz, and C is capacitance in μF .

The inductive reactance, X_L in ohms, is given by:

$$X_L = 2\pi fL \quad (6)$$

where L is the inductance in μH .

At resonance, setting X_c equal to X_L , an expression is found for the reso-

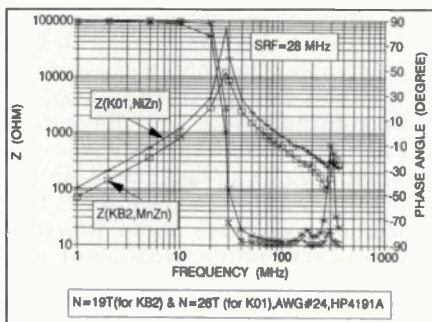


Figure 8. Comparison of impedances of two core materials, nickel-zinc and manganese-zinc.

nance frequency, f_r in MHz:

$$f_r = 1/[2\pi(LC)^{1/2}]10^3 \quad (7)$$

From (7), one can see that the resonant frequency increases as L and C decrease. As L is proportional to N^2 and C is directly proportional to N , the resonant frequency is adjustable by adding or subtracting turns. A 25 turn coil on a nickel-zinc ferrite rod of 0.875 inch length and 0.250 inch diameter, results in an inductance of 15 μH and a self capacitance of 2 pF, giving a

SRF of about 30 MHz.

There may be instances where the use of a differential mode choke is required in order to control conducted emissions in the low kHz to 30 MHz frequency range. At low frequencies, high inductance is needed to obtain adequate impedance, but at the same time effective permeability must be limited due to the necessity of avoiding saturating conditions. High permeability materials such as those used for common mode chokes, saturate at relatively low applied fields, and therefore cannot be used for this purpose. Powdered iron, with its low relative permeability and high saturation magnetization, is very suitable as a core material for low frequency differential mode chokes. Figure 6 is a picture of a toroidal wound powdered iron differential mode choke.

Figure 7 indicates the impedance and phase angle for the same choke measured over the frequency range of 100 kHz to 30 MHz. The SRF is located at approximately 1.5 MHz, thus maintaining a suitably high impedance at both the low and high end of the conducted noise spectrum.

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■ Choke Design

Core Material Considerations

High permeability manganese-zinc ferrite is the best choice as a core material for common mode chokes used in the control of conducted low frequency noise, where the material's permeability allows it to impart a high impedance. The lower saturation flux density of this class of material is not of concern due to the fact that there is no net DC current in the core.

In the case of differential mode chokes for high frequency radiated noise control, a relatively low permeability nickel-zinc ferrite is most suitable. As the optimum design for a differential mode choke is a coil wound on a rod, that is, an open magnetic structure, there is no great advantage in using a material of high permeability to achieve a desired inductance. Use of a nickel-zinc ferrite offers the advantage of high volume resistivity, which allows winding directly onto the core without developing shorted turns. Perhaps the most striking difference between ferrite materials in differential choke applications is the difference in their dielectric constants. The much lower dielectric constant of nickel-zinc ferrite results in a significantly lower self capacitance of a choke wound upon it, compared to a choke wound on a manganese-zinc rod. From (5), capacitive reactance is inversely proportional to the capacitance, therefore, the lower self capacitance of a choke based on a nickel-zinc rod results in a higher impedance at frequencies beyond SRF, compared to one fabricated with a manganese-zinc rod.

Figure 8 illustrates the result of the lower dielectric constant of nickel-zinc ferrite. It shows the impedance versus frequency for two chokes with the same size coil but using different materials. One choke utilizes MMG K01 material, a NiZn ferrite, and the other is made with MMG KB2 material, a MnZn composition. The chokes were designed to have the same SRF. Note that the performance is decidedly different both above and below the SRF, in that the K01 choke has higher impedance over the entire measured band. Above the SRF, this is due to its lower c , as discussed above. Below the SRF, the higher impedance is due to the higher inductance of this choke as a result of its significantly greater number of turns. The high dielectric constant of the manganese-zinc ferrite necessitates a lower inductance to achieve a given SRF.

One advantage of manganese-zinc ferrites over nickel-zinc bodies is their higher saturation flux density. As a

means of comparison, MMG K01 material has a saturation level of 2800 gauss at 25° C, whereas MMG KB2 material saturates at 4800 gauss at 25° C. The maximum DC current a choke can withstand (neglecting wire size considerations) before it saturates is directly proportional to the material's saturation magnetization. Therefore, the choke made with KB2 material is capable of handling about 70 percent more DC current.

Summary

Common mode chokes made with high permeability manganese-zinc ferrite toroids and differential chokes made with powdered iron toroidal cores can be used to eliminate common mode and differential mode conducted noise generated by DC motors. Differential mode chokes utilizing nickel-zinc ferrite slugs have the desired properties for radiated emissions generated by the motor. □

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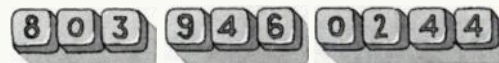
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Lorenzo Carbonini holds a Ph.D. in physics and a Ph.D. in mathematics from the Università degli Studi, Turin, Italy. He has been with Alenia Avionics and Special Equipments Division since 1989. His main technical interest is in the following areas of EMC: devices and antennas for EMC radiated immunity and emission measurements; measuring systems and antenna design for high power radiated immunity measurements on large systems; measuring systems and antenna design for electromagnetic pulse radiated testing.

Mr. Carbonini can be reached at Alenia D.A.A.S., Strada Privata, 10072 Caselle Torinese (TO), Italy. Phone 39-11-9967805.

The subject of this article is the theoretical and experimental analysis of a multiconductor transmission line for radiated immunity measurements. The Horizontal polarization MultiConductor (HMC) antenna (shown in Figure 1) generates both electromagnetic (10 kHz - 200 MHz) and magnetic (up to 1 MHz) fields. The coil for magnetic field measurements is easily created by modifying the antenna geometry. The antenna is very efficient in generating high level fields in a large test region (1.8 m × 1.8 m × 3 m); it is well-suited for performing measurements according to MIL-STD-461D, tests RS101 and RS103, and to IEC 801-3. Further work is in progress to extend the bandwidth of the HMC up to 1 GHz.

Introduction

Some technical constraints are important in determining the feasibility of measuring systems for radiated immunity at low frequencies (below 30 MHz). The main problem, especially for measurements in shielded rooms, is that standard radiating antennas exhibit a low efficiency if smaller than $\lambda/2$, where λ is the wavelength in free space. Hence such antennas must be very large at low frequencies.

Some commercial antennas, based on a parallel plate philosophy, allow measurements at low frequency and are based on a radiating principle. However, the electromagnetic (EM) fields obtainable with such antennas decay rapidly as the distance from the antenna increases. This is due to the fact that the specified test region is in the near field. This is allowed by some regulations such as the MIL-STD-461D.¹

Strong non-uniformities in EM field strength are usually related to an impedance relationship between electric and magnetic field different from that of free space ($|E|/|H| \sim 120\pi \text{ ohm}$). Although equipment under test (EUT) usually operate in such non-uniform EM fields, performing measurements in this kind of field is unacceptable because of the questionable repeatability of measurements done in such fields. To reflect this fact, the IEC 801-3 standard² recommends a minimum distance between antenna and EUT of 3 meters, and a criterium has been introduced to estimate the field uniformity in the test region.

This article describes the features of an antenna for radiated immunity measurements to EM fields in the frequency band 10 kHz-200 MHz (Figure 2). The HMC antenna generates horizontally polarized electric fields. It also performs

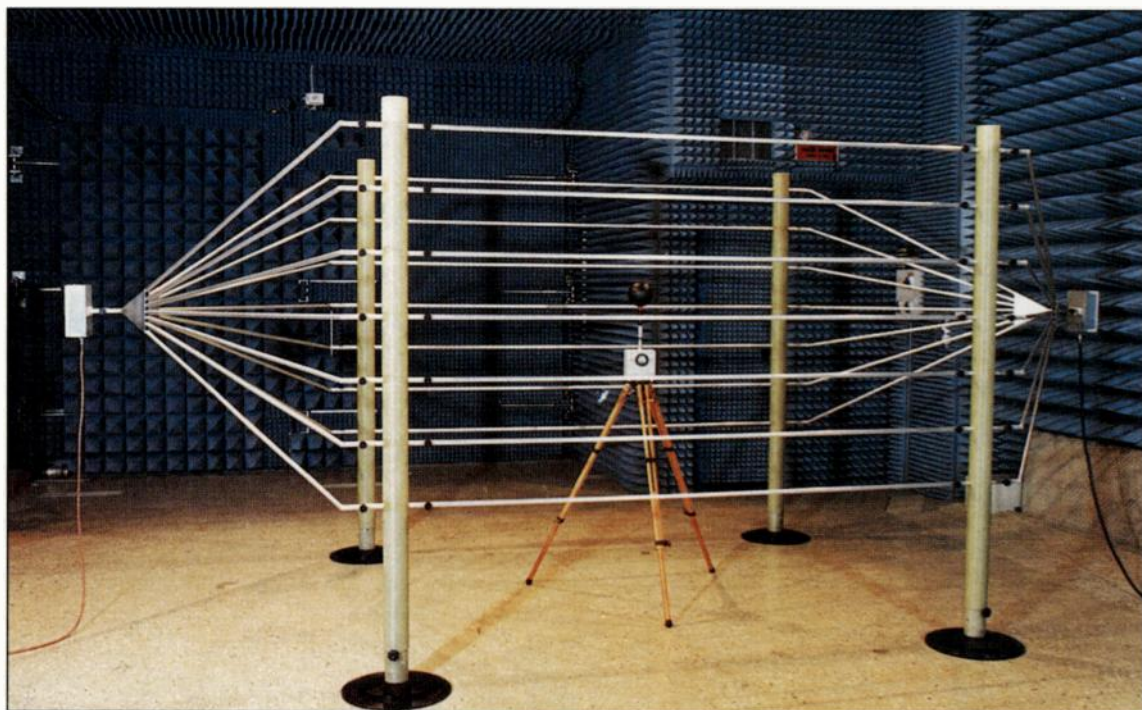


Figure 1. The HMC antenna.

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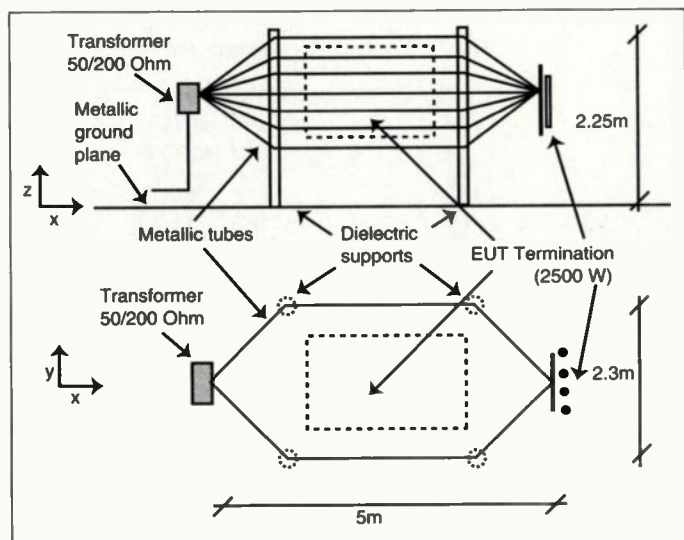


Figure 2. Schematic of the HMC antenna, EM configuration.

magnetic field immunity measurements up to 1 MHz, by turning the HMC into a coil with slight mechanical modifications (Figure 3). Also, low frequency high level electric fields (up to 50 kV/m) can be generated by removing the input transformer and the termination.

The antenna in the EM configuration is based on the principle of transverse EM (TEM) mode propagation inside a transmission line, and is similar to that of classic TEM cells.³ The field distribution at low frequencies is essentially that of the TEM propagating mode. The impedance relationship between the electric and magnetic field for TEM propagation is the same as in free space ($|E|/|H| \sim 120\pi \text{ ohm}$); moreover, the field uniformity for the structure, as will be shown in the following sections, is within $\pm 3 \text{ dB}$, so that an affordable and repeatable testing may be performed by the HMC antenna. A numerical model based on the Numerical Electromagnetic Code (NEC2)⁴ has been used for the theoretical analysis at high frequencies. The results obtained have been also compared with measurements.

Design of the HMC Antenna

The multiwire transmission lines are an efficient solution to perform radiated immunity measurements over a wide frequency band, both in free space and in shielded rooms. In references 5, 6 and 7 the concept has been applied to measurements in shielded rooms, while in reference 8 the application to the case of the WTEM cell is described. It is worth noticing that in reference 8, by experimental results, it is demonstrated that the spurious cou-

pling between multiwire lines and EUTs is substantially lower than for standard TEM cells.³ This feature allows large EUTs to be located inside multiwire lines, yet confirms that the test conditions are very similar to those obtainable in free space, and hence repeatable and comparable to

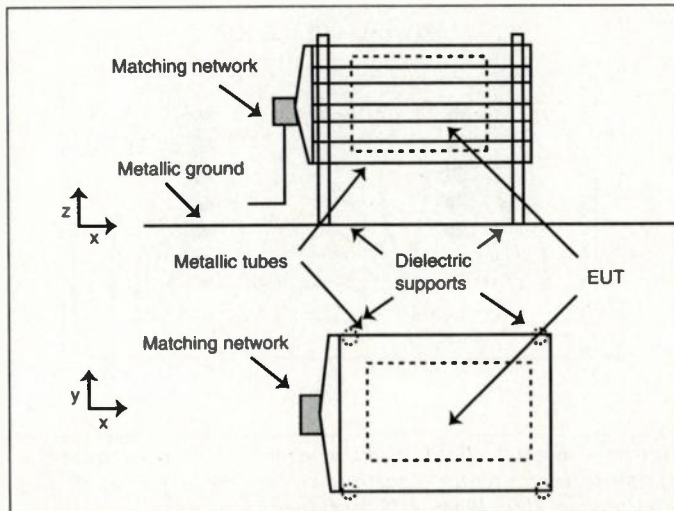


Figure 3. Schematic of the HMC antenna, magnetic configuration.

other solutions.

The HMC antenna design is based on the theory and computer codes for the analysis of multiconductor transmission lines developed in reference 9. A transmission line with 200 ohm characteristic impedance has been designed. Moreover two wideband im-



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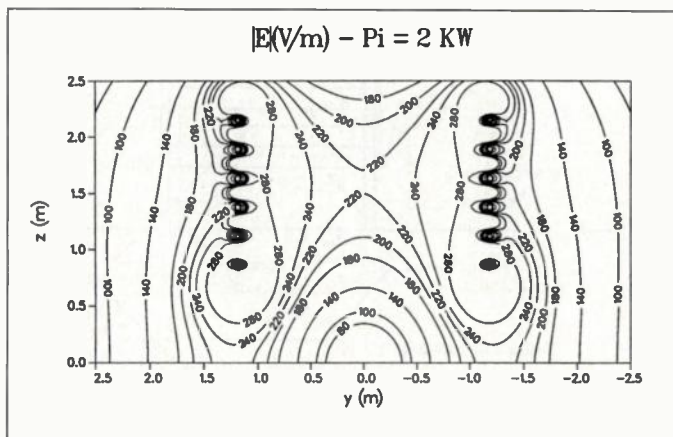


Figure 4. Electric field (V/m) contour plot in a transverse section of the antenna (constant x), low frequency modeling, input power 2000 Watt, EM configuration.

pedance transformers have been developed to ensure a proper matching between the antenna and the amplifying system. The two-dimensional model allows a field prediction accuracy within ± 1 dB when compared with experimental results at low frequencies (below 20 MHz).

Figure 4 shows a contour plot of the simulated electric field obtainable at low frequencies inside the antenna, with a 2000 Watt input power in a transverse section of the HMC line in EM configuration (see Figure 2). The low frequency magnetic field for the HMC in magnetic configuration (see Figure 3) has been computed by a three-dimensional model, based on the quasi-static limit of the relevant equations. A matching network was necessary in order to properly match the antenna to achieve high input current level and proper amplifier operation.

The main advantage of the HMC in

magnetic configuration over conventional Helmholtz coils is that the inductance of the antenna is quite low, allowing easier wideband matching with the amplifier system, especially at high frequencies. Figure 5 shows the magnetic field contour plot obtained in the center transverse section of the antenna with a 1 ampere input current.

Antenna Analysis by NEC2 and Experimental Results

The HMC antenna has been simulated by the computer program NEC2 in the frequency band 1 MHz - 200 MHz, using a double precision version of the code for a better solution stability, especially at low frequencies. The computed voltage standing wave ratio (VSWR) by NEC2 is lower than 1.6 (it assumes an ideal transformer). The measured VSWR is lower than 2 over the whole frequency range, indicating a sufficient impedance match has

been obtained.

Figure 6 reports the electric field level obtainable inside the line with a 2000 watt incident power at the output from the tapered section ($x = 1$ m), as obtained by NEC2. Figure 7 shows a comparison between the field values simulated by NEC2 and those measured at the center of the line ($x = 2.5$ m). The measurements have been performed at full power, in a semi-anechoic chamber with 90 cm pyramidal carbon loaded absorbers. The agreement is generally within ± 2 dB, except for those frequencies at which resonances of the chamber appear (around 28 MHz).

The results show that it is possible to obtain a 200 V/m electric field level over the whole frequency range at the output of the tapered region ($x = 1$ m). Thus, the antenna is well-suited for performing tests according to MIL-STD-461D test RS103.¹

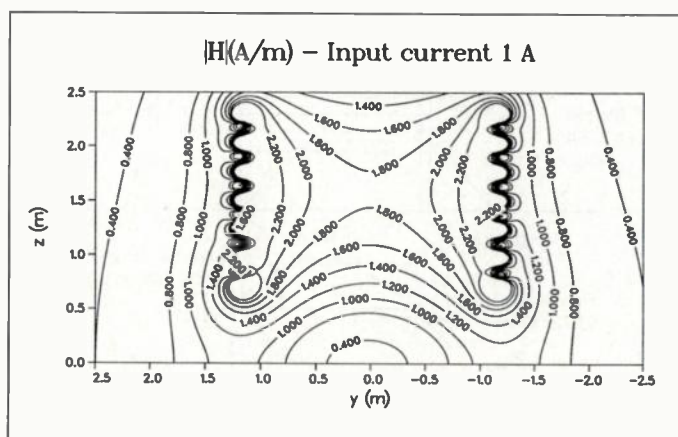


Figure 5. Magnetic field (A/m) contour plot in the central transverse section of the antenna ($x = 2.5$ m), input current 1 A, magnetic configuration.

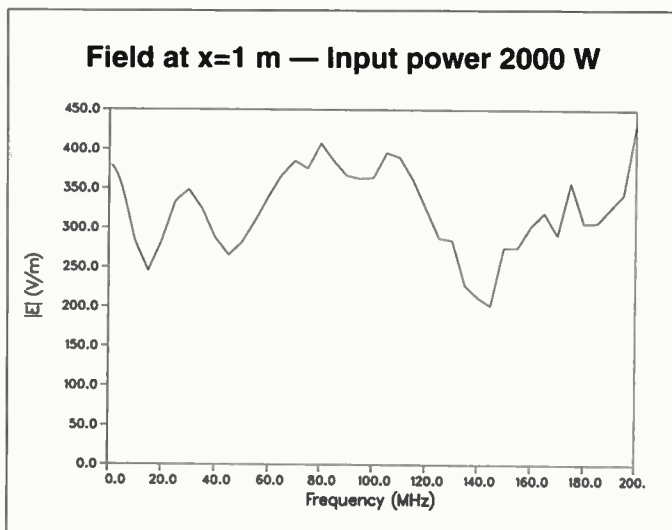


Figure 6. Electric field at the output of the tapered section ($x = 1$ m), simulation by NEC2, input power 2000 Watt.

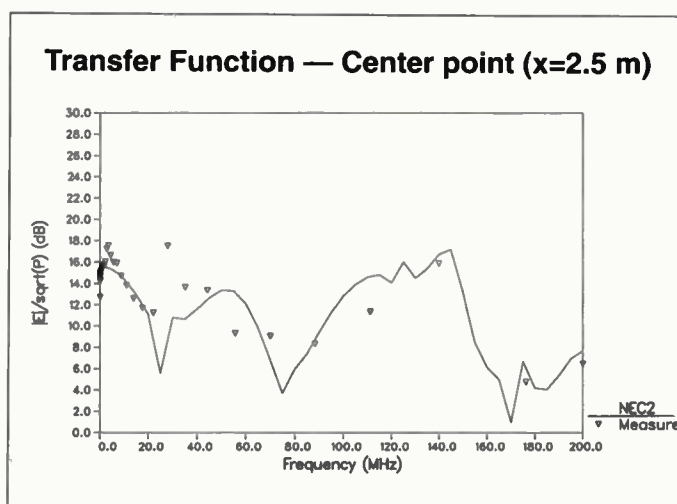


Figure 7. Comparison between the simulated (NEC2) and measured transfer function $|E|/\sqrt{P}$ at the center of the antenna ($x = 2.5$ m).

■ HMC Antenna

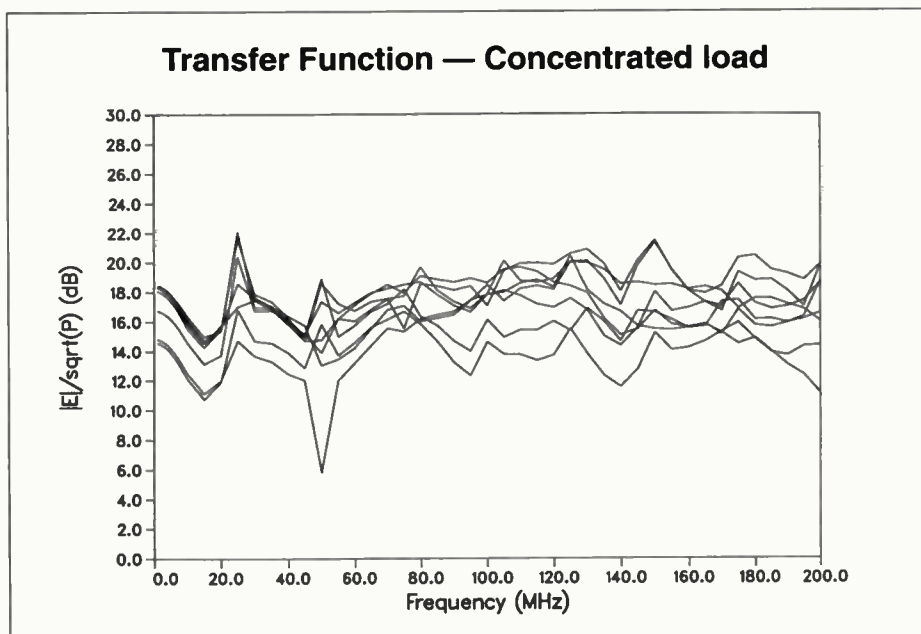


Figure 8. Transfer functions $|E|/\sqrt{P}$ in the points required by the IEC 801-3, HMC in EM configuration with concentrated load, $x = 1$ m.

Moving from the tapering along the line, the fields decrease as the frequency increases. This is due to the traveling-wave antenna behavior at high frequencies, which leads to radiation of power outside the line, and has been already observed in similar antennas.⁵ Further simulations by NEC2 have been performed in order to check the field uniformity, especially at high frequencies, and to check for compliance with regulations such as IEC 801-3.² The field has been computed

on sixteen points spanning a $1.5 \text{ m} \times 1.5 \text{ m}$ surface poised at the end of the tapered transition ($x = 1 \text{ m}$). The results of the simulation are shown in Figure 8. It is worth noticing that there are discrete frequencies (such as 25 MHz and 50 MHz) at which the field values decrease and spread more widely at the same time. This fact is due to resonances of the structure, which create low level standing waves disturbing the field distribution in the test region.

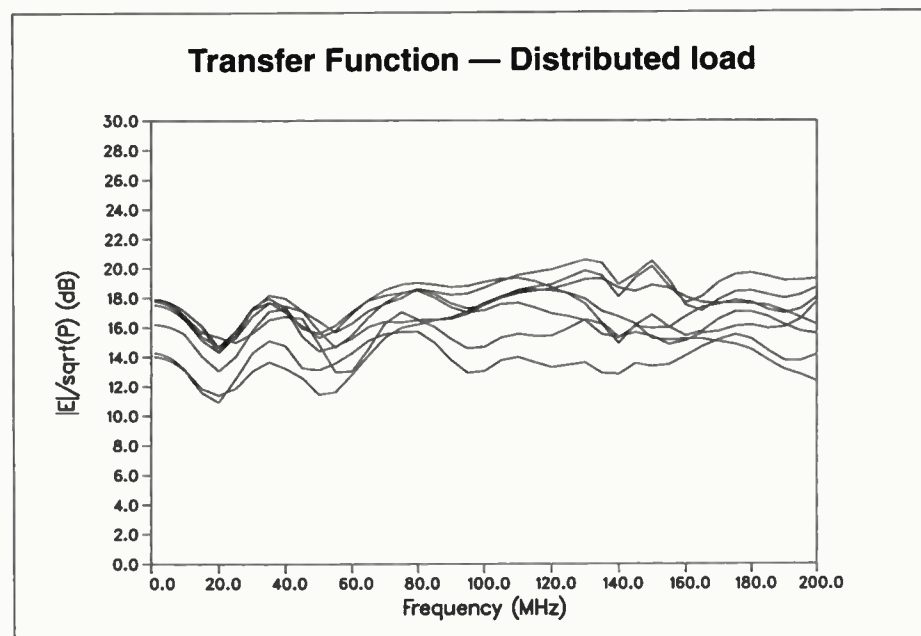
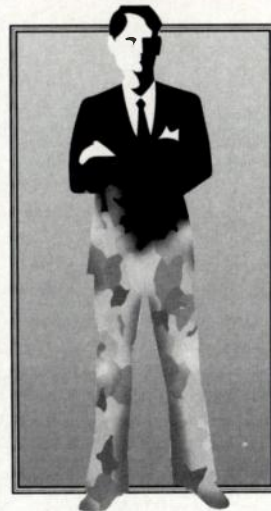


Figure 9. Transfer functions $|E|/|P|$ in the points required by the IEC 801-3, HMC in EM configurations with distributed load, $x=1$ m.

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■ HMC Antenna

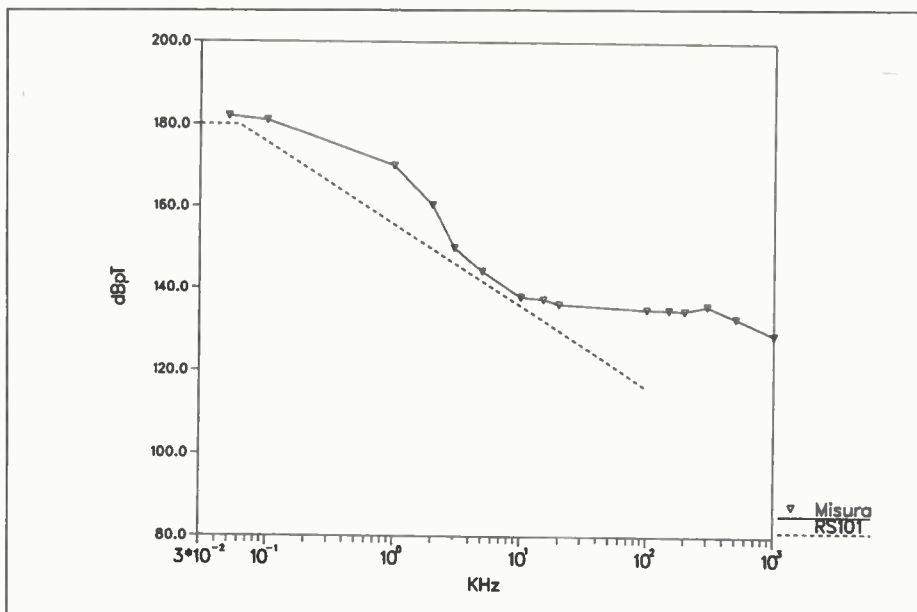


Figure 10. Measured magnetic field and comparison with the limits required by the MIL-STD-461D.

An optimization by software simulation has been performed, aimed at reducing the resonances of the antenna. The standing waves are mainly due to

reflections from the antenna termination, which in the present version is a concentrated load. Improvements may be obtained passing from a concentrat-

ed load to a distributed one, a solution successfully used in EMP simulators and other wideband structures.^{10,11} The results of simulations with a distributed termination are presented in Figure 9. It is evident that the resonances at critical frequencies have been reduced, so that the 0/+6 dB requirement of reference 2 is met over the whole frequency range.

Magnetic field measurements have been performed on the HMC antenna in magnetic configuration (Figure 3). In Figure 10 some experimental results are reported for measurements at the center of the test region, and compared with the limits of the MIL-STD test RS101.¹ The fields shown are obtained by using a low cost amplifier covering the band 20 Hz - 20 kHz, 2 ohm output impedance, 600 watts output power for the low frequency range, and a 2000 watt, 50 ohm amplifier for the high frequency range.

It is worth noting that the HMC antenna, both in EM and magnetic configuration, has field uniformity of ± 3 dB. Though this level of uniformity is not specified in MIL-STD-461D, it does suggest that measurements to that standard will be more repeatable.

For application to IEC 801-3,² the HMC antenna is a good solution to achieve a field level of 10 V/m with 10 watt from very low frequencies, and with the field uniformity required by the regulation. Studies are in progress to redesign the HMC antenna for achieving compliance to the IEC 801-3 up to 1 GHz with low level input power.

Conclusion

The simulations and experimental results confirm that the HMC antenna in EM configuration (Figure 2) is well suited for generating 200 V/m electric fields with 2000 watts incident power in the frequency range 10 kHz - 200 MHz, and meets the requirements of MIL-STD-461D, RS103¹ on large objects 1.8 m \times 1.8 m \times 3 m. Moreover, with some modification of the termination, the antenna is well suited for performing tests according to the IEC 801-3² with field uniformity compliance with low incident power (10 watts).

The HMC antenna, in magnetic configuration (Figure 3), is well suited to perform measurements according to the MIL-STD-461D, RS101,¹ with a commercial, low cost, 600 watt audio amplifier and a 2000 watt higher frequency amplifier (present in many laboratories involved with MIL-STD ra-

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■ HMC Antenna

diated testing). It has been verified that a distributed termination is well suited for improving the field uniformity in the test region. \square

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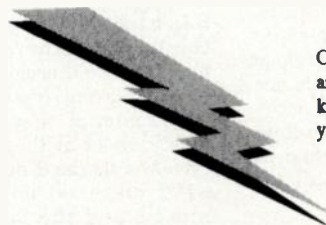
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The Model 30W1000M7 solid-state RF power amplifier from Amplifier Research delivers 30 watts of CW power across a frequency range of 25 MHz to 1 GHz. The Model 30W1000M7 is designed for susceptibility and other RF testing applications requiring level, consistent output power over the test bandwidth. A front-panel gain control with a range of 10 dB minimum allows the operator to adjust power level and radiated

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Credence Technologies, Inc. introduces ScanEM®-H model CTM022, a hand-held, self-contained magnetic-field EMI probe. This probe is a companion to model CTM020, the electric-field EMI probe. ScanEM-H detects the magnetic component of an electromagnetic field and allows engineers to identify the sources of emissions. The probes are available individually or as a kit. Each probe is 5 3/4" x 1 1/4" x 3/4", weighs about 2 oz and can be carried in a shirt pocket. **Credence Technologies, Inc.**
Santa Cruz, CA
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3M Electrical Specialties Division
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Transparent Coatings

IVC has introduced a transparent coating that can be used to provide electromagnetic shielding for window areas of membrane switch panels. The company's ITO transparent coating can be deposited directly using the IVINOX™ process onto screen printed transparent materials used in the manufacture of membrane switches and instrument panels. The shielding coating is normally deposited on the smooth rear surface after screen printing but before adhesive lamination. The normal level of sheet resistivity used is 20 Ω/square.

Inco Vacuum Coatings
West Bromwich, UK
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Foil Laminate Solutions

Tecknit announces a complete line of foil laminates for commercial EMI/RFI shielding applications. Shielding materials include copper, tinned-copper, and aluminum foils laminated to a wide-range of substrates: PVC, Mylar, Kapton, and Nomex. Tecknit will laminate both simple and complex configurations to meet custom design requirements. These materials protect printed circuit boards and plastic enclosures against electromagnetic energy.

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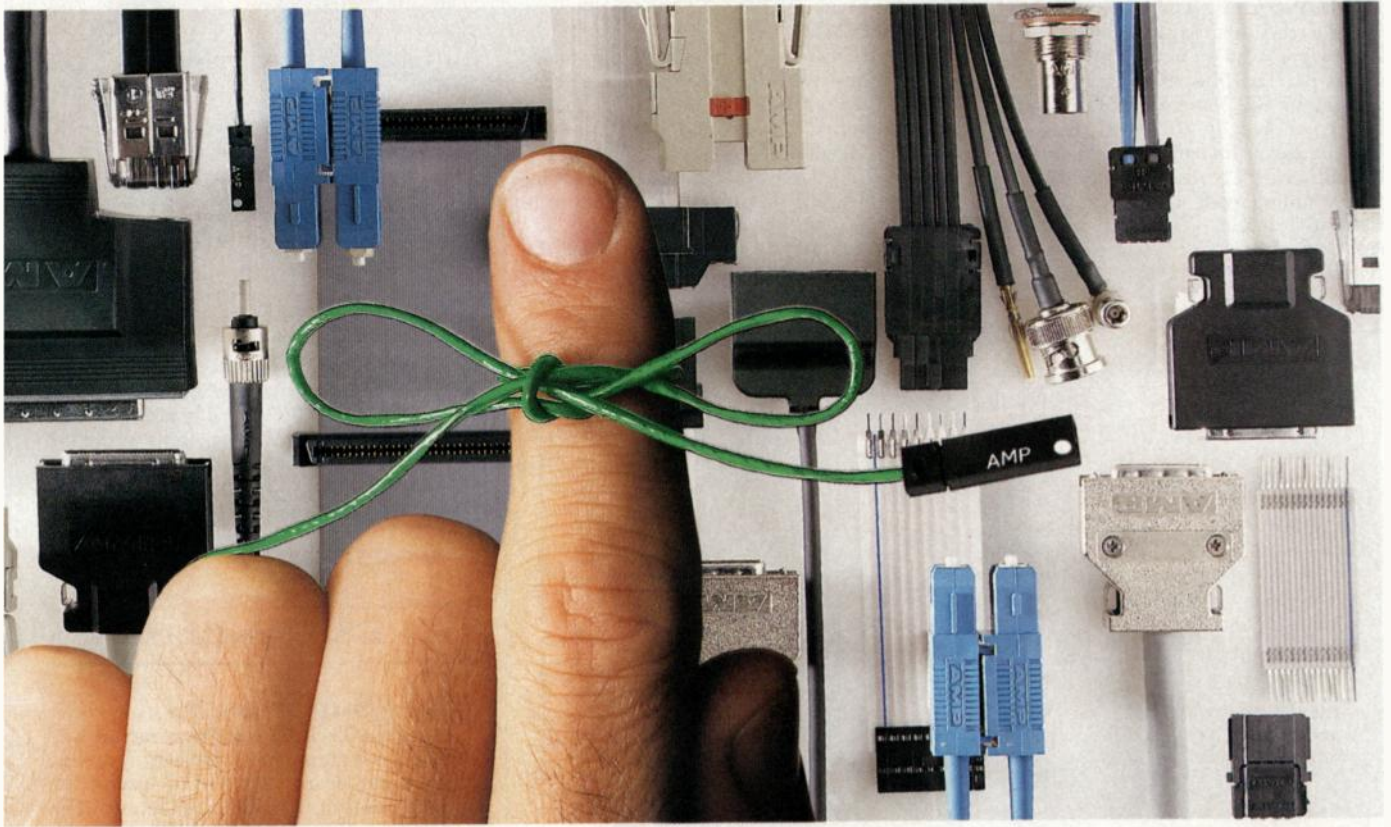
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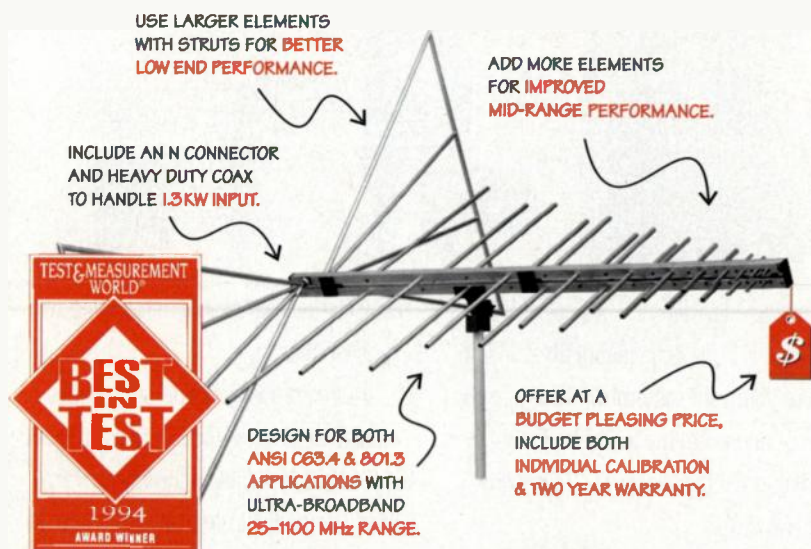
A ferrite filtered connector from Spectrum Control provides filtering of high frequency interference. The ferrite connector provides both pin-to-ground and pin-to-pin filtering. Spectrum's connector features a space saving 0.318" footprint and is interchangeable with standard "D" subminiature connectors. The ferrite connectors address a frequency range of 10 to 300 MHz and provide impedance of 30 ohms at 25



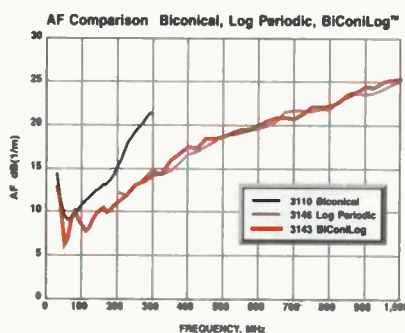
MHz and 50 ohms at 100 MHz. They feature a current rating of 5 Amps, dielectric strength of AC 1000 V for one minute, insulator resistance of 1000 megohms minimum at 500 VDC, and a temperature range of -40° to 105 °C.

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ESD Products

Surge Suppressors

Peradata Technology Corporation has announced its family of multi-port data line surge suppressors designed for protection of data communication equipment. The models SP4, SP8, SP12 and SP16 protect 4, 8, 12 and 16 lines respectively. The multi-port suppressors connect directly between data cables and the equipment to be protected. Each model features a proprietary, three-stage suppression network, including 2100 watt rated suppression device for each data line. The circuitry is packaged in a steel enclosure to shield data signals from outside interference.

Peradata Technology Corporation
Lake Grove, NY
INFO/CARD #157

Series Mode Surge Protector


Zero Surge has developed an OEM version of its series mode surge protector that can be customized to meet designers' specifications. Zero Surge series mode surge protectors reduce the intensity of the surge energy, storing the energy in electrolytic capacitors and then draining it slowly onto the neutral wire without harm to motherboards and other interconnected circuitry. The Zero Surge OEM surge protection device has a master switch, thermal reset button, and 6' line cord. Zero Surge products offer a 10-year limited warranty and lifetime service contract.

Zero Surge
Frenchtown, NJ
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Steward's line of low-profile ferrite components for connectors suppresses EMI before it can couple to cables that serve as troublesome EMI antennas.

Steward's two-terminal lead ferrites and multi-line through hole ferrites offers filtering of low frequency input/output signals entering/exiting shielded enclosures.

Steward's cylindrical EMI suppression ferrites provide a cost-effective means of reducing EMI on the internal and external cable assemblies of electronic equipment.

Steward offers a selection of "split" components for retrofit and post-assembly operations. The rectangular ferrite provides a means of reducing common mode EMI on flat, ribbon type cable assemblies.

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■ New Products

ESD protection products from G&H Technology is designed to be used with standard and high density D-subminiature and a wide range of other I/O connector types. The PULSE-GUARD array can be provided for users who want to add ESD protection to their systems or to connector manufacturers who wish to offer their customers this feature built-in to the connector. The 0.015" flexible PULSE-GUARD array can be installed in seconds simply pressing the array's contacts onto pins at the front or rear of the connector.

G&H Technology
Camarillo, CA
INFO/CARD #155

Literature

EMI/RFI Tutorial

A free tutorial covering the basic concepts of EMI/RFI problems in electronic enclosures and industry solutions is offered by Equipito Electronics. It will be presented by the com-

pany's factory trained representatives at any site requested via a 14-minute VHS tape and product sample. Seminar subjects will include FCC and military specifications. EMP and TEMPEST concerns are covered in detail and alternatives to solving these problems are discussed.

Equipito Electronics Corporation
Aurora, IL
INFO/CARD #154

EMI Shielding Products

Tech-Etch has produced a 20-page catalog on EMI shielding products for doors, panels, covers, connectors, computers, electronic enclosures, and cabinets. The catalog provides general information and product identification on a wide variety of shielding including knitted mesh, thin sheet and strip gaskets, filters, honeycomb vents, and new Quiet Vents.

Tech-Etch, Inc.
Plymouth, MA
INFO/CARD #153

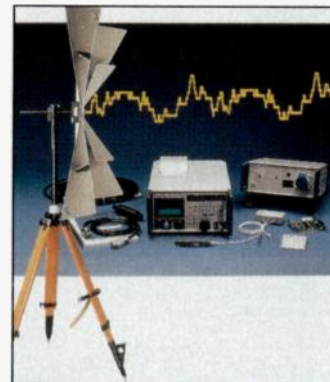
RFDesigner Solution Catalog

A 12-page catalog from ingSOFT Ltd presents the RFDesigner® Solution for RF and microwave design engineers. Packed with detailed technical specifications, the catalog describes the family of software products from ingSOFT Ltd. and third parties, all working together as an integrated design environment. RFDesigner tools provide numerical and graphical analyses, synthesis and optimization of all standard RF components with built-in parasitics and electrical and physical structures.

ingSOFT Limited
Ontario, Canada
INFO/CARD #152

Precompliance Test System

Farnell/Wayne Kerr has released a 26-page literature collection detailing the components of their EASY1 Emissions Assessment System for pre-compliance EMC testing.



Included is data on the SSA1000A spectrum analyzer, broadband antenna, LISN, and EASY1 software. A demo disk for the turnkey system is also available. The turnkey PC based system is aimed at the user who wants to be certain his equipment will meet the stringent requirements of both radiated and conducted emissions for the US and European market.

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Woburn, MA
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De-Certification of PCs and Peripherals Proposed by FCC

By Terry G. Mahn
Fish & Richardson



Terry G. Mahn is a partner in Fish & Richardson. His legal specialty involves international product standards development, harmonization and compliance in the areas of EMC, electrical safety, radio immunity, telephony, bio-effects, radiation and medical safety. He holds a B.S. in Science Engineering from the University of Michigan and a J.D. from Catholic University. Mr. Mahn is a Senior Member of the IEEE and is a IEC CISPR/B committee member and U.S. delegate. He can be reached at: Fish & Richardson, 601 13th Street, N.W., Washington, DC 20005; (202) 783-5070.

Through ET Docket 95-19, the FCC is proposing to eliminate the certification requirements for personal computers and their peripherals. The FCC's objective is to reduce the compliance burden on manufacturers by lowering the costs associated with EMC testing, and speeding up the time it takes to bring new products to market. A summary of the proposed rule changes are set forth in this article.

Declaration of Conformity

In place of a certification requirement the FCC is proposing to allow manufacturers to issue a Declaration of Conformity (DoC) for their PC products. The DoC would have the following requirements:

- Identification of the specific product by trade name and model number
- A statement of Part 15 compliance
- Identification of a compliance test report by date and number
- Identification of the party in the U.S. responsible for ensuring compliance

Under the FCC's proposal, the party *issuing* the DoC would be the party *responsible* for ensuring compliance with all applicable FCC requirements: The DoC would have to be executed before the product could be imported.

Part 15 Product Identifier — The FCC is proposing that all PCs and peripherals display a logo or identifier to signify Part 15 compliance, but is open to suggestions as to what that logo should be. One proposal is to use "North American Class A" (or Class B) as the identifying compliance mark.

Test Laboratory Accreditation — Independent laboratories performing EMC testing for Part 15 compliance would have to be "accredited." One possibility is to use NIST's NVLAP accreditation program, but the FCC is open to other possibilities. For manufacturer's laboratories, the FCC believes accreditation should be required, but is seeking industry comment. Foreign labs would have to be accredited under NVLAP via NIST-recognized accreditation agencies.

Two-Year Transition to DoC Process — The FCC recognizes that laboratory accreditation will take time and proposes a two year transition during which certification of PCs and peripherals will continue to be permitted. Industry input is solicited on whether two years is sufficient.

Increased Sampling by FCC — To ensure that the industry maintains compliance with Part 15,

the FCC is planning to step up its sampling and investigation of devices on the market. EMC test reports would have to be furnished to the FCC within 14 days upon request.

Alternative Approaches Considered — The FCC is inviting comments on alternative approaches to the proposed DoC program. Examples given by the FCC are:

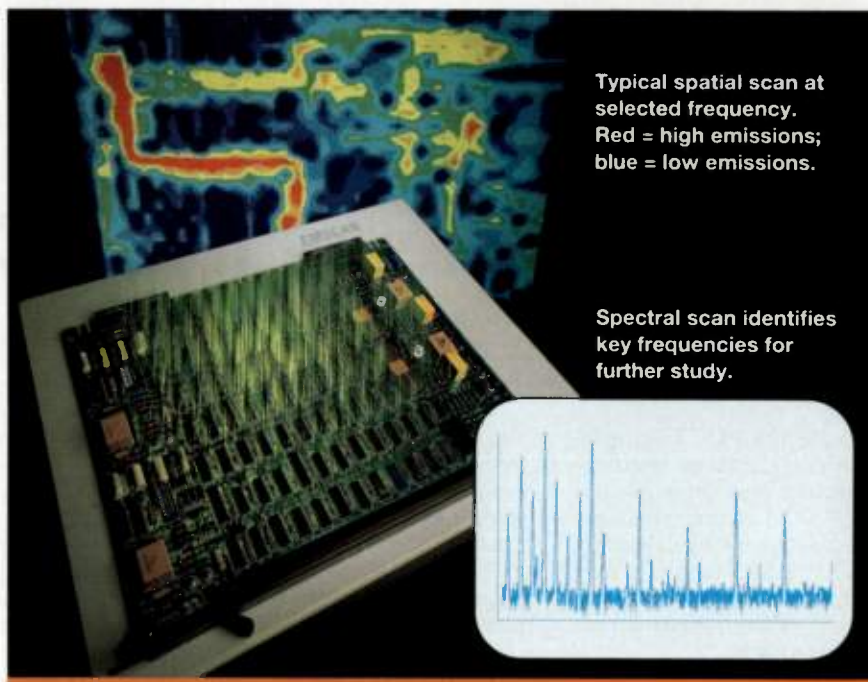
- Retain certification but streamline it to reduce processing time
- Retain certification but permit marketing as soon as an application is filed
- Reduce certification to notification or verification (i.e., no product DoC and no lab accreditation)
- Apply the DoC program to *all* digital devices (e.g., Class A)
- Permit compliance without requiring testing

Modular Component Authorization

An important aspect of the FCC's DoC proposal is to *require* Part 15 authorization for (1) CPU boards, (2) power supplies, and (3) enclosures, that are *marketed to the public*. PCs tested as a system would not be required to obtain separate authorizations for these components unless they are marketed to end-users in component form.

CPU Boards — the FCC classifies a CPU board as a circuit board that contains a microprocessor or frequency-determining circuitry for a microprocessor, the primary function of which is to execute user-provided programming. However, a CPU board would not include a circuit board that contains only a microprocessor intended to operate under the primary control of a microprocessor external to such a circuit or a circuit board that is a dedicated controller for a storage or I/O device.

Authorization for a CPU board would require *two* EMC tests: the first would involve only a radiated emissions test in which the CPU board is connected to a power supply with the oscillator circuit operating and connected to the microprocessor circuit (no peripherals required); and the second would involve both radiated and conducted testing with the CPU board installed in a representative enclosure and configured in a minimum system (see Rule 15.31(a)(6)). Compliance under the first test would be achieved if the CPU board is *within* 6 dB of the Part 15 limits; compliance under the second test must show the emissions to be within the Part 15 limits. The FCC has requested comments on how to treat CPU boards that can accept multiple processors.



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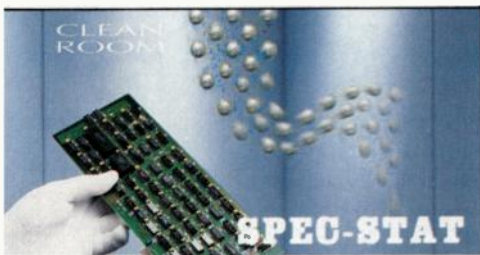
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Power Supplies — Compliance for power supplies would be established by testing them in a "typical" configuration.

Enclosures — Authorization of a PC enclosure would require that it be shown to have a minimum of 6 dB of shielding across the spectrum from 30 MHz to 1000 MHz. In addition, the DoC issued for an enclosure must specify the CPU boards for which it is authorized.

Interchanging Authorized Components — Any party would be permitted to interchange authorized components without the need to retest, provided that a new DoC is issued for the re-assembled system. The new DoC would have to:

- Indicate the basis for compliance (e.g., "only authorized components used in assembly," or "authorized components were installed in an authorized system.")
- Identify each product used in the system
- State that the system complies with Part 15
- Identify the compliance reports for each product by date and number
- Provide the name, address and telephone number of the assembler who becomes responsible for ensuring compliance

Labeling and User Identification — All labeling and user information requirements set forth in the Part 15 Rules would have to be followed for authorized components (see Rules 15.19, 15.21, 15.27 and 15.105). Special accessories for authorized components would be allowed provided they do not involve complex operations

(e.g., soldering or rewiring). The instruction manual for any authorized component would have to specify the installation procedures to ensure compliance, including the "type of enclosure" that must be used.

Marketing of Non-Authorized Components — Non-authorized modular components (i.e., CPUs, power supplies and enclosures) could not be marketed to end-users. OEM sales, however, would be permitted as would sales in "limited quantities" to assemblers for test and evaluation purposes (i.e., beta testing). For importations of non-authorized components, the FCC is requiring that the consignee be the manufacturer/assembler who will assemble and be responsible for testing and authorizing the computer into which the components will be installed.

Editor's note — *The FCC's proposal appears to be a response to various concerns in the PC industry. It addresses the lack of certification of systems assembled by independent shops which select components from many vendors. It also places responsibility for imported components on the importer, a readily identifiable entity. The proposal also removes the requirement of testing all possible combination of components as separate systems, as long as each component is authorized. However, as Mr. Mahn notes, the FCC wants industry input on many details of implementation. The Notice of Proposed Rule Making (NPRM) was issued February 7, 1995. At the time this article was prepared for publication (early March), the NPRM had not yet been published in the Federal Register. Comments will be due 75 days following its publication.*

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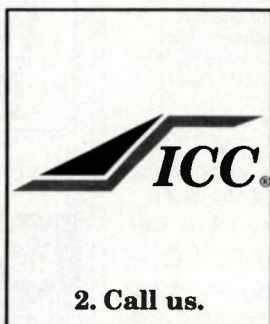
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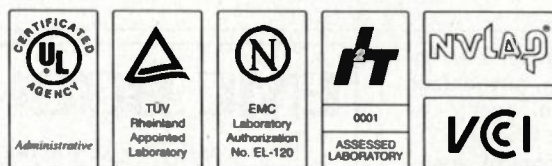


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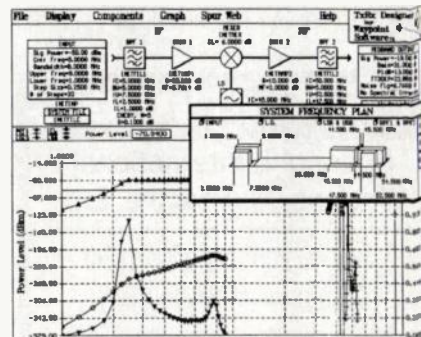
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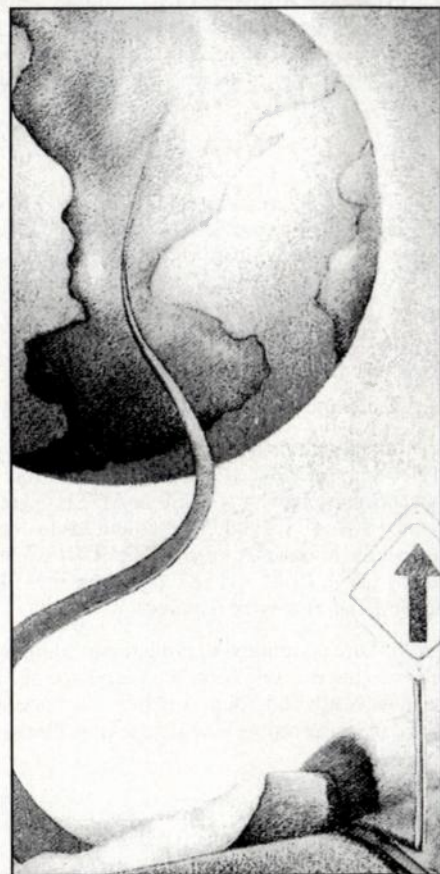
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RF Engineers: Responsible for the design of 800 MHz and 1.9 GHz transmitters and receivers for cellular applications including AMPS, CDMA, and TDMA. Responsible for circuit system specification and analysis of portable products.

Systems Design Engineer: Requires B.S.E.E. (M.S.E.E. preferred) and 5+ years experience. Individual will be responsible for subsystem analysis of baseband for RF signal path for cellular basestation. Design will require an understanding of wireless standards (GSM, DECT, IS-54, PHS, etc.)



Design Engineer Communications ICs: Requires B.S.E.E. (M.S.E.E. preferred) and 5+ years experience. Individual will be responsible for leading the design and characterization of high frequency transceiver ICs for wireless communications applications. Design includes circuit integration of baseband, converter and RF/IF circuitry.

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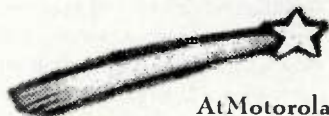
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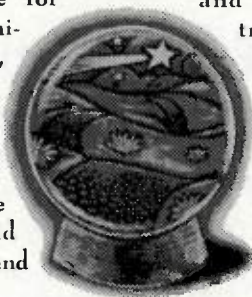
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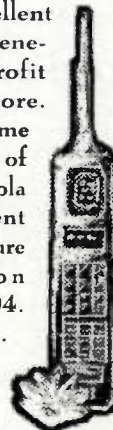
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- **CIRCUIT DEVELOPMENT**—Requires 5+ years' experience defining RF circuitry with demonstrated knowledge of RF concepts such as amplification, oscillation, noise (both phase and thermal), mixing, modulation, filtering, dynamic range, impedance matching and distortion. Strong background in designing, simulating and testing discrete RF circuitry also necessary. (Code BW/RFC)
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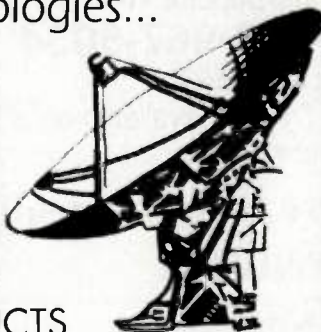
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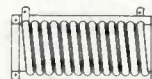
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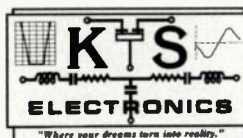
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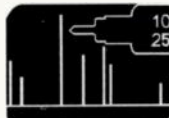
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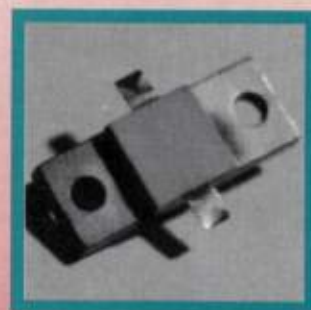
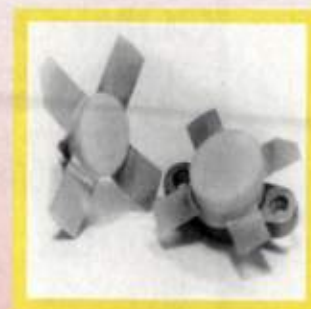
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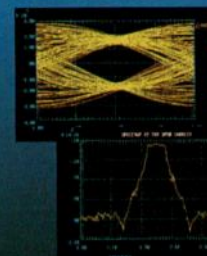
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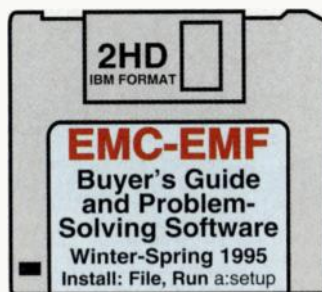
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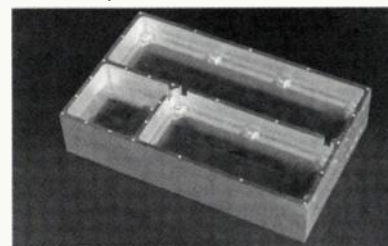
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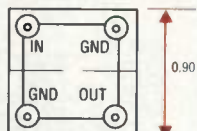


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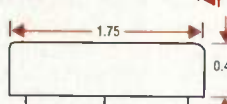
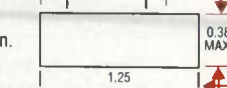
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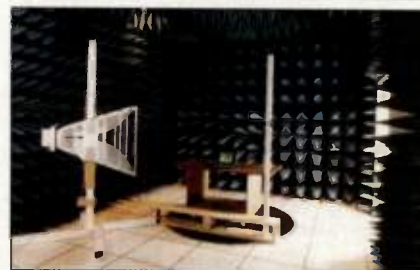
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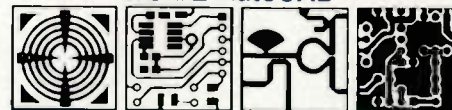
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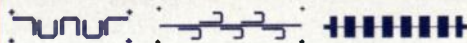
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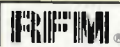
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RF software

Updated Software Suite

HP EEsof has announced the release of version 6.0 of Series IV. This newest version of the UNIX®-based high-frequency electronic design automation suite continues the integration of the Microwave Design System and Series IV products. New features for system analysis include: GMSK/DQPSK coherent demodulators, multirate discrete-time simulation and an antenna and propagation elements option. New circuit analysis features include: planar EM simulator (Momentum), Root FET model, matching network and filter synthesis program. HP Series IV 6.0 is priced from \$16,000.

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Upgraded EMI/EMC Software

Ansoft has upgraded its EMI/EMC and signal integrity analysis software, SI Eminence™. Version 3.2 of the software adds features that allow EMC design engineers to simulate emissions, susceptibility and ESD effects of different design alternatives such as enclosure size and shape, ground structure and placement of material and source characteristics. The upgraded version of SI Eminence is priced at \$49,900 for PC's running Windows NT and workstations running UNIX.

Ansoft Corp.
INFO/CARD #221

3D EM Simulator

Compact Software has begun customer shipping of its version 3.0 Microwave Explorer 3D planar electromagnetic simulator. Microwave Explorer simulates both open and closed environments and runs on Sun SPARCstations and HP 700 series workstations.

Compact Software, Inc.
INFO/CARD #220

Circuit Synthesis

ONYX-CS is a system of stand alone circuit synthesis program modules for the design of microstrip and lumped element circuits. The programs can write DXF™ files, circuit files for a number of circuit analysis programs, and a schematic entry file for ORCAD™. Microstrip circuits include automatic compensation of circuit discontinuities. There are 21 modules. Price is \$490 to \$1795, depending on the number of modules ordered.

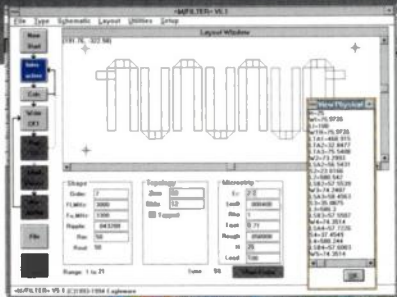
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BER Simulation

The FM software tool now derives bit error rate in digital communications systems using carrier to noise ratio and the SSB phase noise characteristics of equipment oscillators. FM runs on a PC and presents results in a comparable format to that of the actual measurement instruments.

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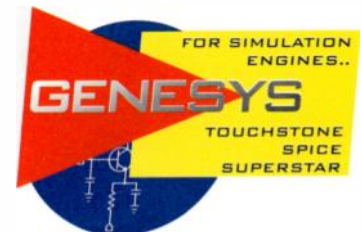
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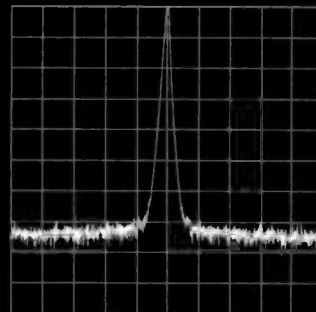


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I_{CC} (typ) @3V	6mA	6mA	6mA	12mA	14mA	15mA	14mA	8mA
Powerdown (typ)	N/A	N/A	30 μ A	30 μ A	30 μ A	1 μ A	1 μ A	1 μ A

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