rf design

engineering principles and practices

August 1989

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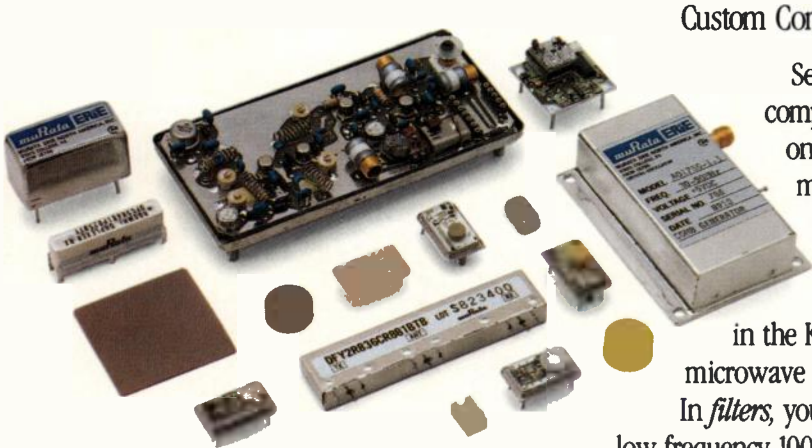

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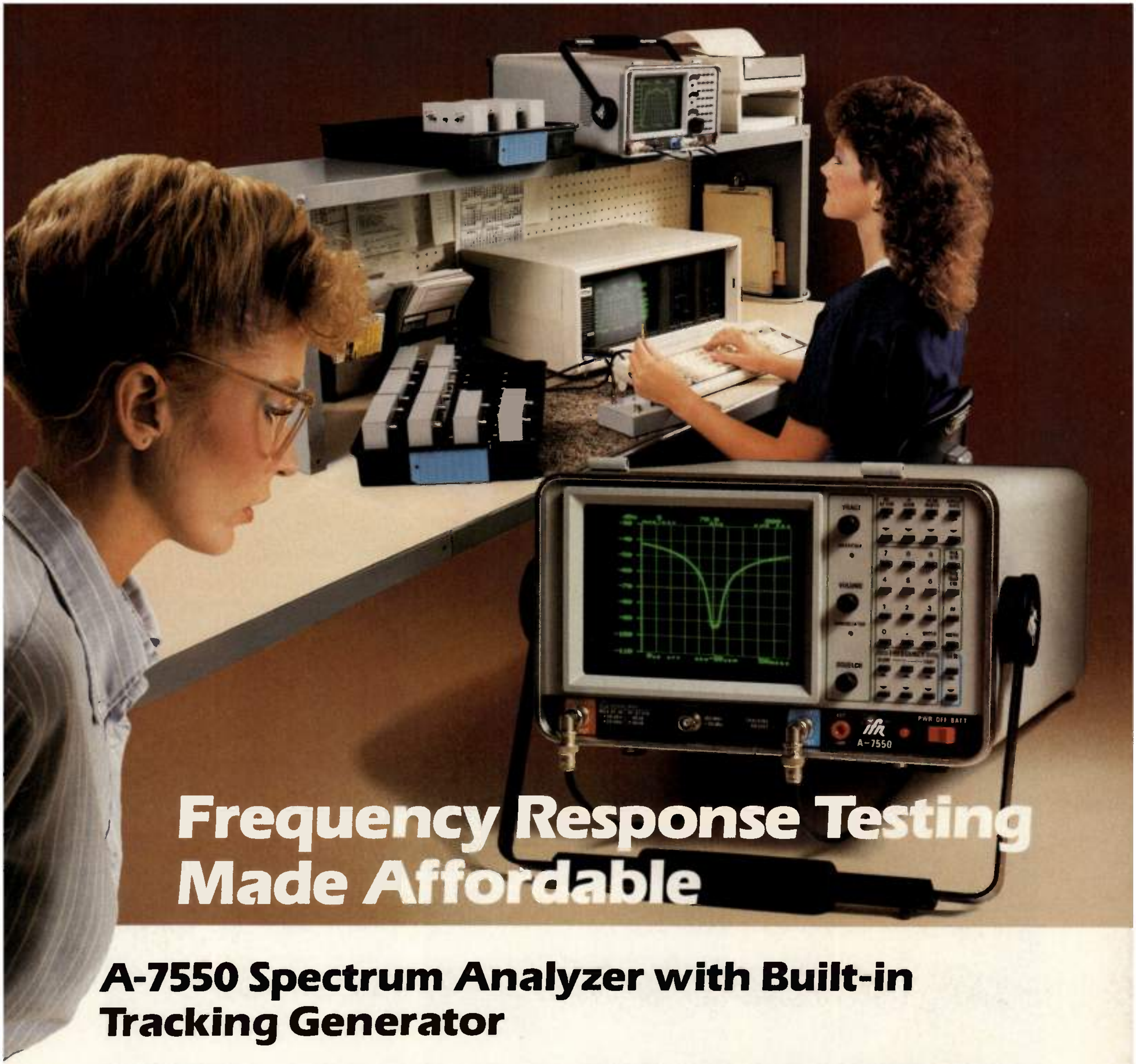

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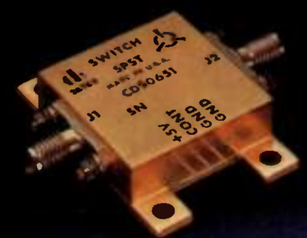

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August 1989

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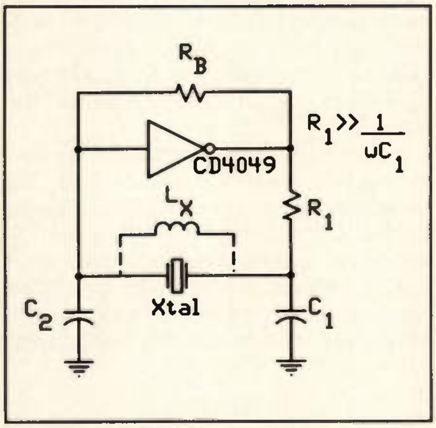
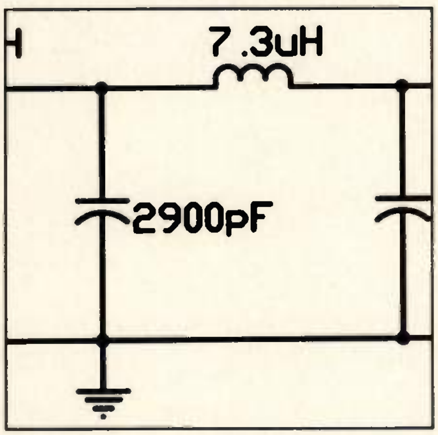


Page 28 — Inverter Oscillators



Page 63 — Chebyshev Filters

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rf editorial

The 1990 RF Design Awards Contest

By Gary A. Breed
Editor

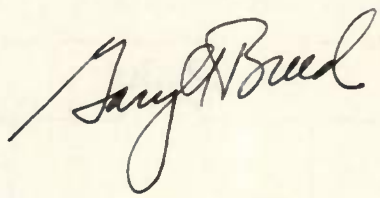
The official announcement of our fifth RF Design Awards Contest can be found on page 36 of this issue. Past contests have begun in October or November, but this year we are starting earlier. Although we receive most entries in the final few weeks before the March 31 deadline, we know they represent work covering several months or more. I think the constant reminder through eight months of advertising will help encourage a few more entries. (So the judges can really be swamped!)

Another reason for announcing the contest so soon is to let you know about our great prizes. The Grand Prize winner will receive a spectrum analyzer, the model R3261A from Advantest. Another engineer will get a precision SMT test fixture and a collection of Design Kits from Coilcraft. Additional Design Kits will be awarded for Honorable Mention entries. We expect to announce even more prizes next month.

Now that we have completed four of them, I am pleased to note that our contest has created far more excitement than we expected. The interest shown by readers who enjoy seeing their colleagues' great ideas is tremendous. RF engineers are a creative bunch, and the chance to see the best ideas of a few of them is inspiring. We will continue to publish contest entries on heavy paper, so you can cut them out and collect them in a notebook.

Another pleasant response to the contest is our advertisers' desire to support it with prizes. For the first few years, the prize providers were enthusiastic, but they understood that the contest was not a guaranteed success. Fortunately, they all found that their support of RF engineering was recognized and appreciated. Continued support from RF companies, like this next collection of prizes, will go a long way to ensure continued success of the contest.

Your circuit might be the next winner, joining Dan Baker's phase detector, his frequency/amplitude calibrator, Charlie Wenzel's frequency multiplier, and Al Helfrick's optically-coupled VCO. Start thinking about it now, then get those ideas to us by the end of next March. Maybe you'll be on our July 1990 cover!



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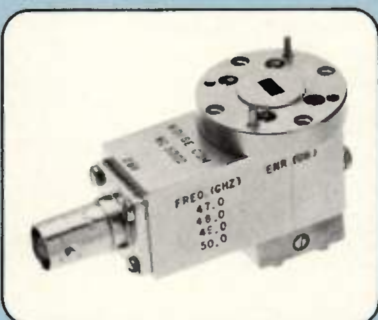

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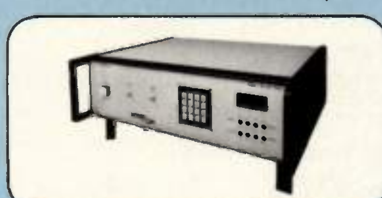

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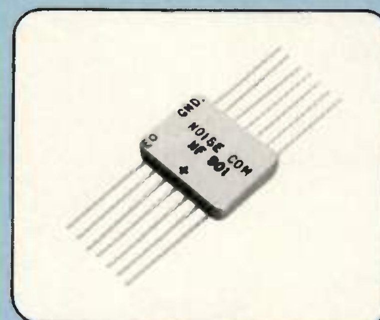

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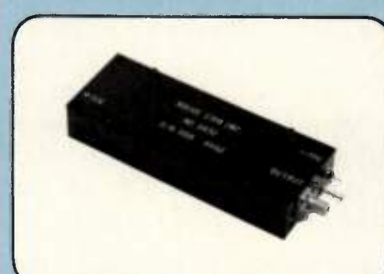

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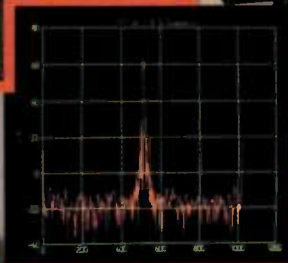

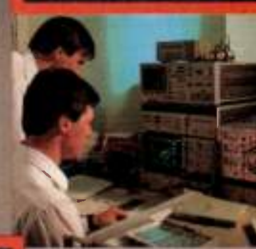

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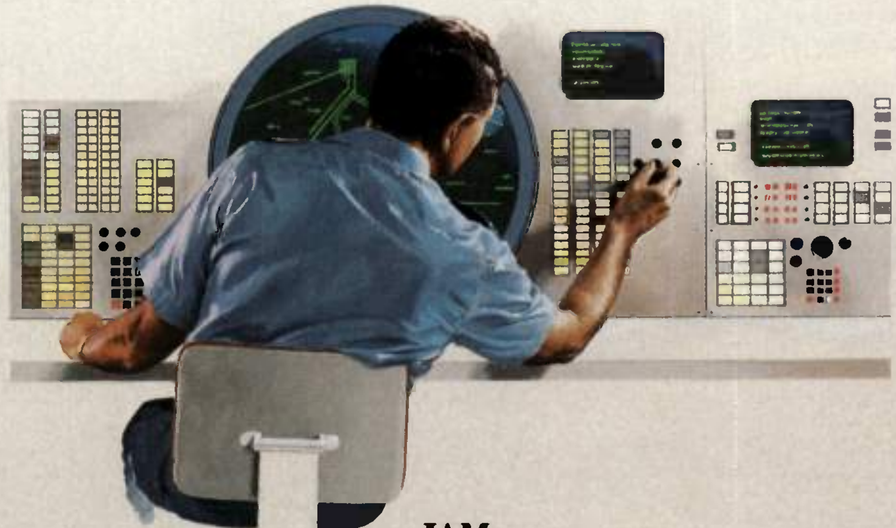

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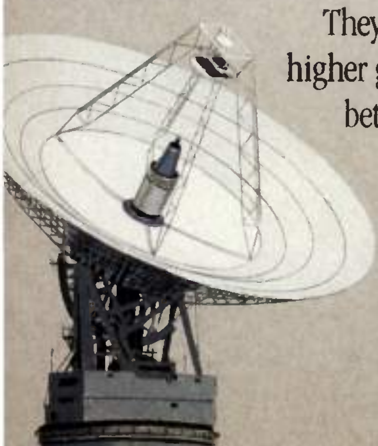

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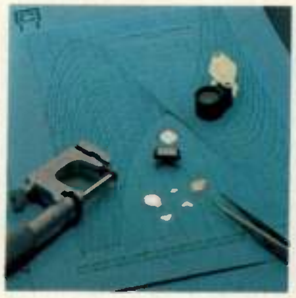


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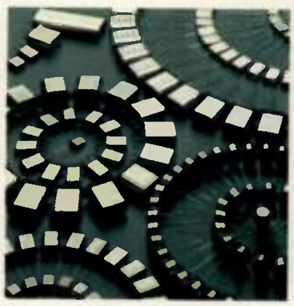


Equatorial

rf letters

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

A Common Error

Editor:

I am writing in reference to the article, "Designing Facilities for Lightning Protection" (*RF Design*, June 1989). The equation

$$V = L \frac{di}{dt}$$

is incorrectly given, which is a common but basic error made by many electronics authors. The correct relation is:

$$V = -L \frac{di}{dt}$$

That little minus sign is not unimportant, as it signifies that a *collapsing* current, denoted by:

$$-\frac{di}{dt}$$

produces a *positive* voltage across the inductor, L. That is a fundamental law of electrophysics.

Paul Galluzzi
Dynamics Research Corporation
Wilmington, Massachusetts

Article Index Request

Editor:

I would consider very useful an annual index of *RF Design* articles, either printed in the last issue of each year, or recorded on diskette. Recording on a disk would allow inclusion of the indexes of five to ten years, arranged by topics, authors, etc.

Carlos M. Bielli
Quark Electronica SRL
Buenos Aires, Argentina

The December 1989 issue will include an index of articles from 1988 and 1989. — Editor

A Well-Wisher

Editor:

I greatly appreciate your initiative in creating the RF Design Software Service. May it be a fruitful idea!

V. Grigoras
S.C.S. Laboratories
Iasi, Romania

An Informed Opinion

Editor:

It's obvious from the articles and other valuable information that *RF Design* is committed to keeping us informed. And I mean up to the minute!

T. Renfro
Butler Service Group
Kokomo, IN

Recent reader suggestions for future articles have included:

- Directional couplers
- DROs, DTOs
- Spread spectrum
- Radiolocation
- Low-frequency quadrature hybrids
- VHF-UHF power amplifiers
- Digital EMI reduction
- More computer programs

Corrections

The following corrections to "A General-Purpose Oscillator" (*RF Design*, June 1989) should be noted.

- In Figure 3, the supply voltage is +12 V.
- In Figure 4, the frequency is 1.62 GHz.
- In Table 2, the phase of S_{11} is 177 degrees.
- The reference on p. 61 to 1.68 GHz should read as 1.62 GHz.
- In Appendix 1, for the first "Z" equation, the Z_{22} part of the matrix should read:

$$(1-S_{11})(1+S_{22}) + S_{12}S_{21}$$

- In Appendix 1, for the "A" matrix ladder network, A_{21} for Z_c should read:

$$j\tan(\beta l_c)/Z_c$$

- In Appendix 1, the denominator for the final "S" equation should read:

$$1/((Z_{11}''+1)(Z_{22}''+1) - Z_{12}''Z_{21}'')$$

- In Appendix 1, the final paragraph should read, "Note that Z_e is the characteristic impedance..."
- In Appendix 1, the diagram for Z_f should be an open circuit.

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This month's disk (RFD-0889)

1. "Chebyshev Filters with Arbitrary Source and Load Resistances," by Jack Porter of Cubic Corp. (BASIC).
2. "A General Purpose Oscillator," by Dr. Y.C. Cheah of Hughes Network Systems. From June 1989 issue. (Mathcad files for S-parameter conversions).

Disk RFD-0789: A collection from 1988 issues

1. "Predicting RF Output from Combined Power Amplifier Modules," by Roderick Blocksome, from February 1988 issue (Lotus 1-2-3™ spreadsheets).
2. "Equal-Ripple LC Filter Synthesis," by Robert Kost, February 1988 issue (compiled, executable code).
3. "Microstrip Analysis and Design of Various Substrates," by D.R. Hertling and R.K. Feeney, June 1988 issue (BASIC).
4. "CAD Amplifier Matching with Microstrip Lines," by Stanley Novak, June 1988 issue (BASIC).

(Add $5.00 for reprints of these four articles.)

Disk RFD-0689 (June 1989)

"A BASIC Program for PLL Design," by James Conn (BASIC).

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Kansas City Westin Crown Center, Kansas City, MO
Information: EIA, Components Group, 1722 Eye Street N.W., Washington, DC 20006. Tel: (202) 457-4930

August 29-31, 1989
International Test Conference '89
Sheraton-Washington Hotel, Washington, DC
Information: ITC '89, P.O. Box 264, Mt. Freedom, NJ 07970. Tel: (201) 895-5260

September 4-8, 1989
19th European Microwave Conference and Exhibition
Wembley Conference Centre, London, England
Information: Microwave Exhibitions and Publishers Ltd., 90 Calverly Road, Tunbridge Wells, Kent TN1 2UN, England. Tel: (0892) 44027; Fax: (0892) 41023

September 8-10, 1989
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Information: EMC'89/Nagoya, Prof. Yasumitsu Miyazaki, Toyohashi University of Technology, 1-1, Aza-Hibarigaoka, Tempaku-cho, Toyohashi-City, Aichi, 440 Japan. Tel: (81) 0532-47-0111 ext. 528; Fax: 0532-45-0480

September 10-14, 1989
9th Annual International Electronics Packaging Conference and Exhibition
Sheraton Hotel, Harbor Island, CA
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September 12-16, 1989
2nd International Conference and Workshop on Electromagnetic Compatibility
Bangalore, India
Information: Col. (Dr.) G.Kl. Deb, Electronics and Radar Development Establishment, CV Raman Nagar, Bangalore 560 093 India.

September 26-28, 1989
International Conference on Lightning and Static Electricity
University of Bath, Surrey, England
Information: Laura Christie, ERA Technology Ltd., Cleeve Road, Leatherhead, Surrey KT22 7SA, England. Tel: 0372 374151, ext. 2290; Fax: 0372 374496

October 24-26, 1989
RF Expo East 89
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Information: Kristen Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-2600; (800) 525-9154

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8×10^{-10} at 1 Sec
Long Term Stability:
$< 1 \times 10^{-6}$/year
Warm Up:
< 20 seconds to $\pm 1 \times 10^{-7}$
Input Voltage: 15 V ±5%
5 V (TTL)
Input Power:
< 0.5 W During Warm-up
0.38 W Stabilized at Room-Temp.
Size: 1.26″ × 1.26″ × 0.7″
Output Waveform:
TTL (optionally Sine)

OCXO
Oven Controlled Crystal Oscillator

Frequency Range:
to 50 MHz
Short Term Stabilities:
up to 5×10^{-12} (1 sec)
Warm-Up Time:
As low as 1 min
Temperature Stability:
$\pm 5 \times 10^{-10}$
(0^0 to $+50^0$C)
Low Aging Rate:
$< 5 \times 10^{-11}$/Day
Low Noise:
< –157 dBc@
10 kHz Offset
Low Vibration Sensitivity:
3×10^{-10}/g
Temperature Range:
-55^0 to $+120^0$C

TCXO
Temperature Controlled Crystal Oscillator

Frequency Range:
0.02 Hz to 20 MHz
Frequency Stability:
±0.8 PPM
(-40^0 to $+85^0$C)
Aging: ±1.0 PPM/yr typ.
Supply Voltage:
2 to 15 Vdc
Supply Current:
As low as 1.0 mA
Size:
Standard:
1.5″ × 1.5″ × 0.5″
As small as:
0.960″ × 0.5″ × 0.2″

CXO
Crystal Clock Oscillator

Frequency Range:
TTL: 10 Hz to 100 MHz
C-MOS: 1 Hz to 5 MHz
ECL: 5 MHz to 500 MHz
Sinewave: 1 Hz-1 GHz
Frequency Stability:
Typ. ±50 PPM (-20^0 to $+70^0$C, Industrial)
Typ. ±50 PPM (-55^0 to $+125^0$C, Military)
Up to ±10 PPM available (-20^0 to $+70^0$C)
Aging:
±10 PPM/yr (Industrial)
±5 PPM/yr (Military)
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rf courses

The George Washington University

Integrating Fiber Optics and Analog/RF
August 21-23, 1989, Washington, DC
Radar Operation and Design: The Fundamentals
August 29-31, 1989, Washington, DC
Microwave Radio Systems
September 7-8, 1989, Washington, DC

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

UCLA Extension

Kalman Filtering
August 28-September 1, 1989, Los Angeles, CA
Non-Gaussian Signal Processing and Applications
September 11-13, 1989, Los Angeles, CA
Analog MOS Integrated Circuits
September 25-29, 1989, Los Angeles, CA

Information: UCLA Extension, P.O. Box 24901, Department K, Los Angeles, CA 90024-0901. Tel: (213) 825-3344

University Consortium for Continuing Education

Sonar Signal Processing
September 18-22, 1989, Washington, DC

Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995-6335

Compliance Engineering

EMI
August 22, 1989, San Jose, CA
Safety
August 23, 1989, San Jose, CA
ESD
August 24, 1989, San Jose, CA
Telecom
August 25, 1989, San Jose, CA

Information: Compliance Engineering, 629 Massachusetts Avenue, Boxboro, MA 01719. Tel: (508) 264-4208

Design and Evaluation, Inc.

Worst Case Circuit Analysis
October 16-18, 1989, San Francisco, CA

Information: Design and Evaluation, Inc., 1451B Chews Landing Road, Laurel Springs, NJ 08021. Tel: (609) 228-3800

EMC Services

Filter Design for Switching Supplies
August 21-22, 1989, Chicago, IL
October 2-3, 1989, Washington, DC
EMI Control for Switching Supplies
August 23-25, 1989, Chicago, IL
October 4-6, 1989, Washington, DC

Information: Sonya Nave, EMC Services, 11833 93rd Avenue North, Seminole, FL 34642. Tel: (813) 397-5854

EEsof, Inc.

Computer-Aided Engineering for Nonlinear Microwave Circuits (LIBRA)
August 21-23, 1989, Westlake Village, CA
Computer-Aided Engineering for Nonlinear Microwave Circuits (mwSpice)
August 24-25, 1989, Westlake Village, CA
System Design (OmniSys)
September 11-12, 1989, Westlake Village, CA
Computer-Aided Engineering/Drafting for Microwave Circuits (Academy)
September 18-20, 1989, Westlake Village, CA

Information: Sande Scoredos, Training Coordinator, EEsof, Inc., 5795 Lindero Canyon Road, Westlake Village, CA 91362. Tel: (818) 991-7530, ext. 197

Hewlett-Packard Co.

Designing for Electromagnetic Compatibility (EMC)
August 28-29, 1989, Naperville, IL
August 31-September 1, 1989, Novi, MI
September 18-19, 1989, North Hollywood, CA
September 21-22, 1989, Fullerton, CA

Information: Hewlett-Packard Co., 3000 Hanover St., Palo Alto, CA 94304. Tel: (800) 2HP-EDUC

Integrated Computer Systems

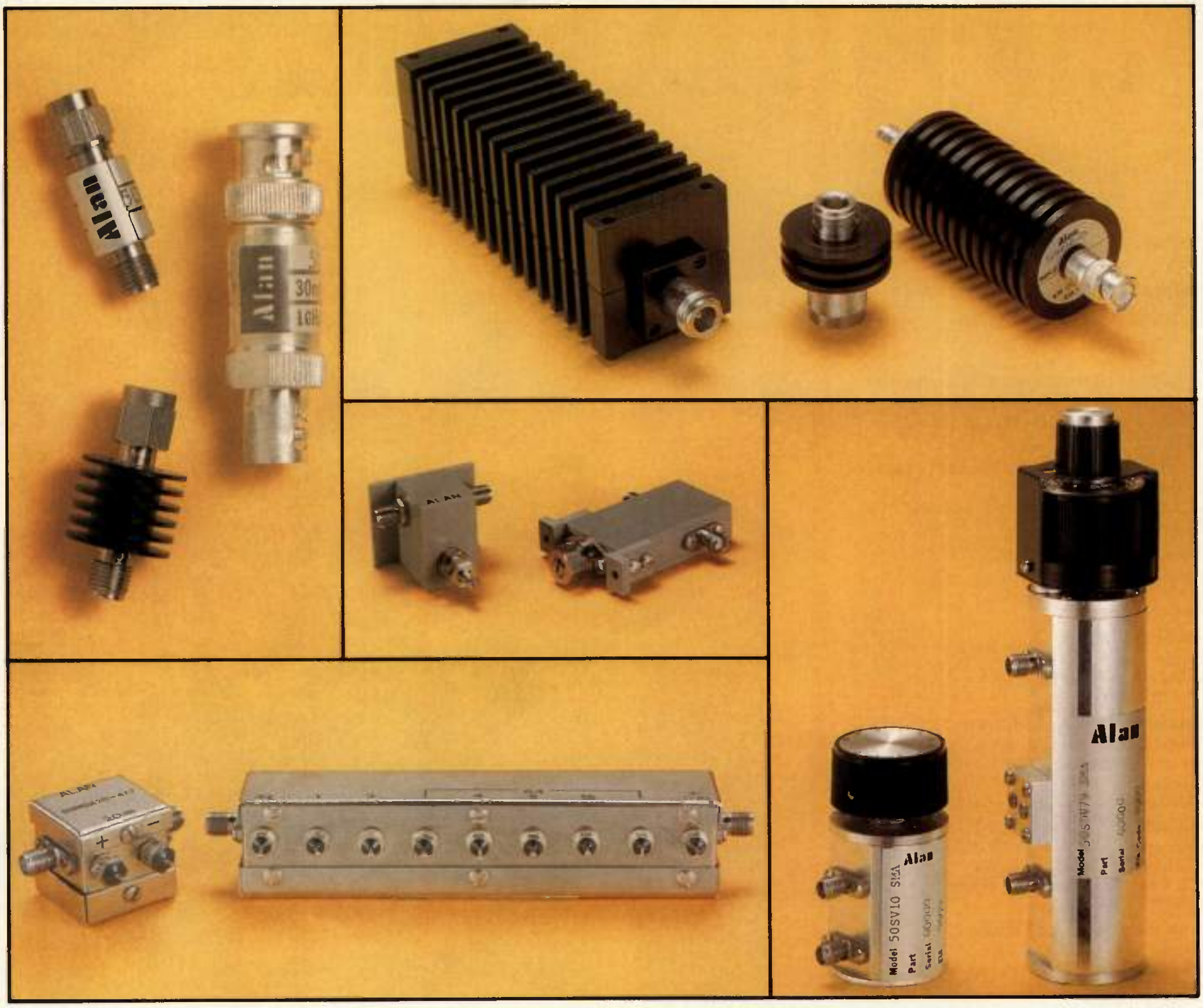
Introduction to Fiber Optic Communications
August 22-25, 1989, Los Angeles, CA
August 29-September 1, 1989, Washington, DC
Introduction to Telecommunications
August 22-25, 1989, San Diego, CA
September 12-15, 1989, Washington, DC
C Programming Hands-On Workshop
August 29-September 1, 1989, Washington, DC
September 12-15, 1989, Denver, CO
Troubleshooting Datacomm and Networks
August 29-September 1, 1989, Washington, DC
September 19-22, 1989, San Francisco, CA

Information: John Valenti, Integrated Computer Systems, 6053 W. Century Boulevard, P.O. Box 45974, Los Angeles, CA 90045-0974. Tel: (800) 421-8166; (213) 417-8888

Interference Control Technologies, Inc.

Grounding and Shielding
August 22-25, 1989, Annapolis, MD
Practical EMI Fixes
September 11-15, 1989, Washington, DC
TEMPEST Facilities Design, Installation and Operation
September 11-15, 1989, San Francisco, CA
High-Speed Digital Design
September 12-15, 1989, San Diego, CA
EMC Design and Measurement
September 18-22, 1989, San Diego, CA
Intro to EMI/RFI/EMC
September 26-28, 1989, Orlando, FL

Information: Penny Caran, Registrar, Interference Control Technologies, Inc., State Route 625, P.O. Box D, Gainesville, VA 22065. Tel: (703) 347-0030



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Volt-time product: 42 - 372 V-µsec
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Inductance: 5 - 300 µH
18 styles (48 total pieces)
Kit P205 $75

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Current: .03-4.3 amps
Inductance: 3.9 µH - 100 mH
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RF Attenuators and Switches: An Update

By Mark Gomez
Technical Editor

As it has been in the past, the demand for RF switches and attenuators is predicted to remain constant. Even with the shrinking military budget, this segment of the RF industry seems to be strong due to the activity in the commercial arena. "The switch market is very strong," observes Nils Musaeus, marketing manager at Vari-L. "There are many new applications springing up," says Vernon Hickson, product specialist at Kay Elemetrics, adding, "I wonder if these commercial applications will supplement the lack of military business." Bill Kennedy, president and general manager of Alan Industries, anticipates market growth. He notes, however, that the companies involved in the military business will see a shrinking market. "The commercial market will step in and take its place," he remarks, going on to say that the Europeans could also have a positive impact on this market.

Joetta Walker, marketing manager at JFW Industries, reports that her company is experiencing some growth this year. "This is because of expanded markets like communications, and new RF industries such as medical electronics," she explains. This viewpoint is shared by Bruce Malcom, president and CEO of Trilithic. "The whole market is increasing, and this is due to the increase in commercial applications."

The anticipated growth is evident in special product areas like switch matrices, GaAs (gallium arsenide) switches and GaAs attenuators. "I see more inquiry for electromechanical single housing switch matrices both in low frequency and microwave frequency applications, especially for automatic test equipment," observes Mel Okano, a staff consultant with Loral Microwave-Wavecom. In switches and attenuators (aside from mechanical switches and fixed attenuators), there seem to be four dominant technologies: relays, Schottky diodes, PIN diodes and GaAs. The newest is GaAs. "GaAs FETs are a growing fad in switches and attenuators," says Bill Grunau, vice-president of marketing and engineering at Daico Industries.

Although constant market activity is forecast, industry experts predict some kind of fallout in the companies serving this market. Due to aggressive competition, smaller companies and companies without strong financial backing may be forced to merge or pursue niche markets in the industry. "There are many companies in the field, and the market seems to be getting tighter," observes Hickson. "Some of the smaller companies, unless they have strong financial backing and are truly aggressive in their sales effort, may experience a fallout," comments Kennedy. Steve Ulett, sales manager at Penstock, feels that there will be fewer companies and some mergers in this industry. "For survival, in some cases, you will see companies gel together," predicts Walker.

Although prices are relatively low for products that have been around for a while, more competitive pricing can be expected due to the proliferation of companies. In areas that are relatively new, prices should stabilize and remain comparably flat. "We will see more competitive pricing in the areas where there is large competition," observes Walker. "The standard product that has been around for a long time is going to get more competitive," she adds. "Both in generic switches and attenuators, especially in the standard lower frequency attenuators, prices can be expected to go way down," remarks Ulett.

This pricing trend is no doubt related to advances in technology. As better and more inexpensive ways of building a product are invented, prices generally fall. "People are figuring little advances that make for relatively major savings," says Hickson. According to Ray Vincent, senior vice-president at Microlab/FXR, people in the business are constantly looking for ways to reduce cost so they can be more competitive. Bruce Malcom observers, "We can expect to see lower cost and better performance switches and attenuators."

Together with lower cost, features such as denser packaging, higher reliability, faster switching speeds, better component characterization, higher frequencies and overall better performance are common requests in the attenuator and switch industries. "Everybody is looking for smaller size, lower cost and higher reliability," says Carl Schraufnagl, vice-president of marketing at KDI/triangle Electronics. According to Jim Andrews, president of Picosecond Pulse Labs, customers are pushing higher and higher frequency limits and are requesting smaller and smaller packages.

The use of GaAs FETs is one of the ways that manufacturers are achieving higher speeds. Nevertheless, this technology does have its drawbacks. "GaAs switches provide fast switching speeds and offer very low power consumption," explains Grunau. "However, they have a relatively higher insertion loss." Another problem in GaAs switches is the settling time. "With GaAs switches, it can take 20 ns to switch from 50 percent TTL to 90 or 95 percent RF and take another 100 to 300 ns to settle to 100 percent," he adds. With GaAs attenuators, problems with voltage-controlled units are also evident. The third-order intercept point at the middle of the attenuation range tends to be unacceptable for most applications where harmonic and spurious performance is important.

The switch and attenuator market seems to be in sound shape. Although there are cutbacks in military applications, the commercial sector seems to be compensating for them. Also, attenuator and switch customers should benefit from the positive changes illustrated by improved specifications as well as lower costs.

Bush Nominates Sikes to FCC Chair—Alfred C. Sikes, director of the Commerce Department's National Telecommunications and Information Administration (NTIA), has been named by President Bush to fill the post of chairman of the Federal Communications Commission (FCC). The post is being vacated by Dennis Patrick, whose resignation was announced last April. Sikes has served as NTIA's director since 1986. Prior to that he was president of Sikes and Associates, a Springfield, Mo., consulting and broadcasting management firm. Sikes has also served as assistant attorney general and director of consumer affairs, regulation and licensing for the state of Missouri.

The Bush administration also announced appointments to fill two vacancies at the Commission. Andrew Barrett, a member of the Illinois Commerce Commission since 1980, and Sherrie Marshall, a Washington, D.C., attorney and former director of the FCC's Office of Legislative Affairs, have been named to fill two vacant Republican commissioner seats. Both nominations, as well as that of Sikes to the chairmanship, are subject to Senate confirmation. Current chairman Patrick's resignation is effective upon Senate confirmation of the new chairman.

The term of Patricia Diaz Dennis, one of two Democrats on the Commission, officially expired at the end of June; she remains on the job, however, pending confirmation of a successor. Commissioner Dennis has publicly expressed interest in being renominated to the Commission. The remaining seat on the five-member commission is held by Democrat James Quello.

DARPA Names Five Firms to Receive HDTV Money—The Defense Advanced Research Projects Agency (DARPA) has named the first of the companies that will be receiving money for development work on high-definition television (HDTV). Five U.S. companies have been selected for funding of research into HDTV display technologies: Raychem Corp., Menlo Park, Calif.; Texas Instruments, Dallas; NewCo Inc., San Jose, Calif.; Projectavision Inc., New York; and Photonics Technology Inc., Northwood, Ohio. The first four companies will be working on projection display technology, while Photonics Technology will be investigating plasma flat-panel display technology.

The contracts were awarded as part of a three-year, $30 million HDTV program announced by DARPA earlier this year. Nearly 90 proposals were received by DARPA in response to its request for proposals. $15 million of the total funding is earmarked for display technologies, with the remaining $15 million to be spent in development of display processor technology. The exact amounts awarded under the first five contracts have not been announced. The awarding of these contracts marks the first outlay of government money to support research work into HDTV.

Report Predicts Strong Growth for U.S. Test Equipment Market—The dramatic rise in prominence of digital communications indicates strong growth ahead for the U.S. test equipment market. A new study from Frost and Sullivan Inc. predicts $840 million a year will soon be spent on test equipment, a result of the increasingly important role communications is playing in industry and government. *Communications Test Equipment Market in the U.S.* notes that, as data exchange intensifies, "advances in digital technology and fiber optics have not only revolutionized the manner in which information is transmitted, but have also been responsible for the major revamping of facilities involving large capital investments." Thus test equipment associated with digital technology, fiber optics and data communications will fare extremely well in the coming years.

The study's overall prediction is that the nearly $551 million in communications test equipment that was sold in 1988 will amount to $840 million (constant dollars) annually by 1993. Data communications test equipment will comprise the largest category throughout the period discussed. The growth rate for this segment is forecast to exceed the market's overall average for the next several years. By 1993, volume in this area will represent 28 percent of all dollars invested in communication test equipment.

The report, A2051, is available at a price of $2,400. For further information, contact Frost and Sullivan Inc., 106 Fulton Street, New York, NY 10038. Tel: (212) 233-1080

Device Offers Non-Destructive Testing—A new instrument, developed at the National Institute of Standards and Technology (NIST), makes possible the non-destructive testing of electronic switching components. The new device is the work of David W. Berning, an electronics engineer at NIST. The instrument, an automated power transistor switching test system, determines the maximum voltage and current levels that the device under test can switch without destroying it.

The new instrument operates on the principle that the temperature rise which actually destroys the transistor, or other power electronic device, occurs slightly after the electrical breakdown of the device. The breakdown occurs when the electric current reaches a point where the device can no longer stand the voltage in the circuit. "That's the actual event we test," explains Berning. "We wait until the voltage collapses. That collapse triggers our instrument, which removes the power." The instrument operates at up to 2,000 V and 100 A. The delay from the time the voltage collapses to the complete removal of current depends on the current and is measured in nanoseconds. Testing to the 40 A level, for example, requires about 25 to 30 nanoseconds. Berning's latest version of the instrument includes automation for repetitive testing, and is the first known self-contained, semi-portable test system that does not require external power supplies, pulse generators, oscilloscopes, and other monitoring devices to make a transistor safe-switching test.

Varian/HP Process Offers Improved GaAs Device Yields— Varian Associates Inc. and Hewlett-Packard Co. (HP) have jointly developed

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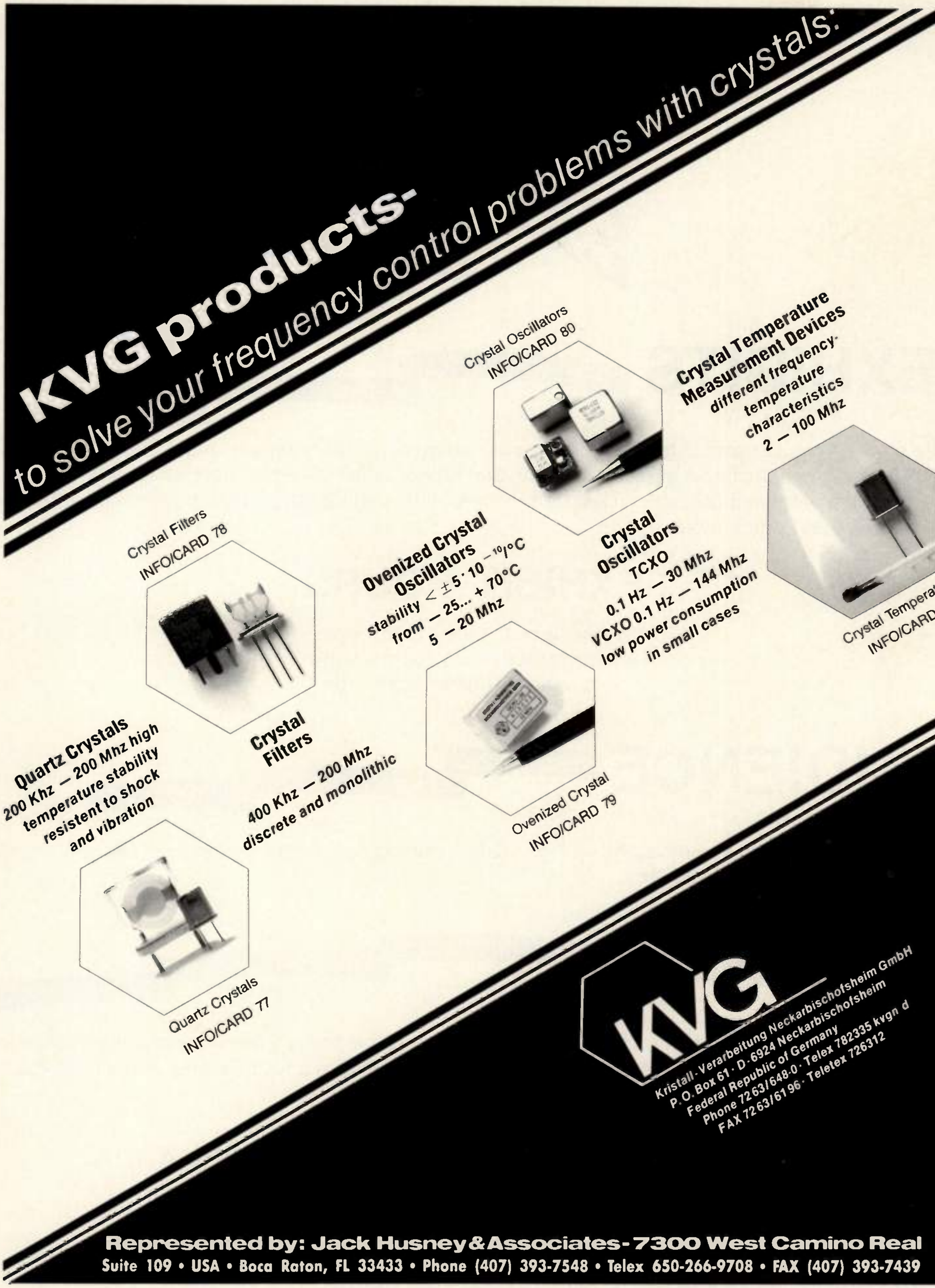
rf news *Continued*

a process for producing gallium arsenide (GaAs) integrated circuits (ICs) that offers the potential for significant improvements in yield over existing manufacturing techniques. The process uses a metal-organic molecular beam epitaxy (MOMBE) system.

Researchers at Varian's III-V Device Center recently fabricated high-electron mobility transistor (HEMT) devices and ICs using GaAs wafers manufactured at HP with a Varian/HP MOMBE system. HP and Varian report that the MOMBE HEMTs have performance characteristics as good as those fabricated using conventional molecular beam epitaxy (MBE) wafers, but have defect densities up to ten times lower than the typical MBE wafers. Low defect densities contribute to higher device and IC yields, which translate into lower costs.

The results are the outcome of a joint development effort begun by Varian and HP in 1987 to develop a MOMBE machine and fabrication process. In addition to military applications addressed with HEMT devices, Varian expects that the MOMBE process will make possible the economical production of high-performance analog receiver circuits anticipated for high-definition television (HDTV).

RF Curing May Improve Carpet Productivity—Researchers at Georgia Tech's School of Textile Engineering are studying a method for curing latex carpet-backing which incorporates RF drying. Results indicate that an RF oven could be used successfully to supplement a traditional drying system for water-removal during the curing of latex adhesive on the back of tufted nylon carpet. Most U.S. carpet and textile manufacturers use conventional convection ovens, which dry from the outside inward, to remove water. For bulkier, heavier products like carpet, however, the conventional process can result in slow and uneven drying, reducing product quality and increasing processing time. RF dryers operate by producing electromagnetic energy which causes certain molecules in the product to oscillate rapidly, generating heat "from the inside-out." Researchers at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VCXO 0.1 Hz — 144 Mhz
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in small cases
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Georgia Tech believe a system employing RF drying/curing in conjunction with a conventional oven would result in more complete and uniform drying of bulky textiles. In addition, the drying time could be reduced from several hours to just a few minutes.

NVLAP Lab Directory Available— *The 1989 Directory of NVLAP Accredited Laboratories* lists 200 labs nationwide and abroad that are accredited by the National Institute of Standards and Technology (NIST) for specific test methods in various fields of testing as of April 1, 1989. Among the fields covered in the 1989 Directory are electromagnetic compatibility and telecommunications, acoustics, computer protocols, and personnel radiation dosimetry. Labs are listed alphabetically by name, field of testing, and state. Copies of the directory are available prepaid for $15.95 from: National Technical Information Service, Springfield, VA 22161. Order by PB #89-189278.

Marconi to Purchase Racal Instrumentation Units—Racal Electronics plc and Marconi Instruments Ltd., a subsidiary of General Electric Co. plc, have reached an agreement according to which Marconi will acquire three Racal businesses: Racal-Dana Instruments Inc., Irvine, Calif.; Racal-Dana Instruments Ltd., Burnham, England; and Racal Automation Ltd., Burnham, England. According to the agreement, Marconi will purchase the three companies for about $48 million in cash. Racal Electronics said the decision to sell followed a critical evaluation of the high research and development investment required to maintain competitiveness in world markets. The units purchased produce RF instruments, automatic test equipment, electronic counters, digital test subsystems, and switching systems.

Harris Wins Contract for Canadian Navy Communications System— Harris Corp.'s Long-Range Radio Division has been awarded a $19.3 million subcontract by Paramax Electronics Inc., of Montreal, to supply a high-frequency (HF) radio communications system for the Royal Canadian Navy. The system includes HF transmitters, receivers and antennas for shipboard applications, including the Canadian Navy's new patrol frigates.

The Harris system represents a technology breakthrough in broadband HF radio system design. HF transmitters and receivers, operating simultaneously in close proximity aboard ships, potentially cause signal interference. The Harris system design solves this problem through an approach that combines multiple radio signals in a single broadband transmitter.

The Harris broadband transmitting system allows multiple exciters to be amplified and fed to a common antenna while reducing most of the intermodulation products to more than 110 dB below the carriers. In contrast to most power-bank type systems, all signals are not run through a common power amplifier. With up to four exciters, the Harris broadband transmitting system has no common amplification and produces third-order intermodulation distortion products more than 70 dB down.

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INFO/CARD 21

Kansas City Hosts Quartz Industry

The 11th Quartz Devices Conference and Exhibition will be held August 29 through 31, 1989. Once again, the Westin Crown Center in Kansas City, Mo., will be hosting this industry gathering. Sponsored by the Electronic Industries Association (EIA), the event will bring device users and manufacturers together for three days of technical presentations and a full exhibit floor.

The conference is organized around two tracks or sessions, focusing on technical and productivity topics, respectively. Session A, Applying Current Technology, will consist of technical papers directed towards those experienced in the frequency control field as manufacturers or users of quartz crystal products.

The following presentations are scheduled as part of Session A:

Tuesday, August 29
9:00 a.m. to 12:00 noon

- X-Ray Fluorescent Thickness Measurement for SPC (Statistical Process Control) of Films on Quartz
- Computer-Aided Failure Analysis for Wideband Crystal Filters
- Field-Assisted Bonding of Single-Crystal Quartz
- High-Reliability Crystal Devices: Planning for Development and Manufacture
- Why Publish?

Tuesday, August 29
2:00 p.m. to 5:00 p.m.

- Steady-State Oscillator Analysis in Immittance Domain: Part 1
- Precision Temperature Test Station for Quartz Crystals
- Plating Crystal to Frequency Using EIA-512 Methods
- Realization of High-Degree Delay Equalizers for Crystal B and Pass Filters
- A Very Fast Determination of the Orientation of Doubly Rotated Quartz Cuts

Wednesday, August 30
9:00 a.m. to 12:00 noon

- Measurement of Quartz Crystal Units Up to 500 MHz and Above by the Use of a Pi-Network With Error Correction
- Influence of Stray Reactance on Measured Crystal Parameters
- Quartz Crystal Resonator Parameter Sensitivity Study
- Abrasive Selection and Control in Lapping Operations
- Report on Activities of IEC/TC-49
- Report of Activities of EIA Engineering Committee P-11
- Report of TC-49 and P-11 Activities on Quartz Materials
- Report of Activities of TC-49 Measurements Subcommittee
- Report on Oklahoma University Work on Quartz Materials

Wednesday, August 30
2:00 p.m. to 5:00 p.m.

- Digitally Compensated Quartz Crystal Oscillator in a Compact Housing
- Steady-State Oscillator Analysis in Immittance Domain, Including Computer-Aided Design: Part 2
- The Influence of Design Parameters on the Dynamic Capacitance of Quartz Crystal Units
- Generating Surface Figure and Finish on Quartz: Part 2

Session B, which runs concurrently with Session A, is devoted to Improving Productivity. It includes sections entitled "How to Market and Advertise Your Company" and "Current and Future EPA and OSHA Rules and Regulations and What They Mean to You." Also scheduled in Session B is a one-and-a-half day workshop, "Designing a World-Class Operation With Manufacturing Cycle Time Reduction."

A general session on Thursday morning, "Future Piezoelectric Device Requirements," will assemble representatives from government and major market segments to discuss processes and applications relative to piezoelectric device usage in the future. The military, telecommunications, automotive and industrial segments of the quartz industry will be among those areas represented in these discussions.

Highlights of this year's program include two awards presentations. The Piezoelectric Devices "Man of the Year" award will be presented during Tuesday's exhibit hours, as will the award for "Best Paper" from the 1988 Quartz Devices Conference. That paper, "Quartz Resonator Model Measurement and Sensitivity Study," reprinted in this issue of *RF Design*, reports on research into an accurate mathematical characterization of quartz crystals, based on extensive measurements.

The exhibit hall will be open Tuesday and Wednesday, August 29-30, from 2:30 p.m. to 7:00 p.m., and Thursday, August 31, from 9:30 a.m. to 12:00 noon. Exhibiting companies include:

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Crystal Components
EG&G Frequency Products Inc.
Electro Dynamics Crystal Corp.
Emkay Manufacturing Inc.
Engineered Materials
Fotofabrication Corp.
GSM Inc./Quartz Devices Directory
H.E.S. International Inc.
Highland Mfg. Co.
Houston Electronics
K.C. Abrasive Co., Inc.
Kay Inc.
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M-tron Industries Inc.
Masden Industries
P.R. Hoffman Materials Processing
P.R. Hoffman Machine Products
Philips
Polaris Electronics
RF Design
Saunders and Associates Inc.
Sawyer Research Products Inc.
Schott Electronics
Shure Cut Crystal
Slee-Superior
Thermo Dynamics
Transat Corp.
Valpey-Fischer Corp.
Voltronics Corp.

Proceeds from the Quartz Devices Conference and Exhibition for the past six years have been contributed toward supporting extensive university/industry directed piezoelectric resonator, oscillator and filter research and experiment programs. Northern Illinois University, Oklahoma State University and the University of Central Florida are among the institutions which have participated in this program. Present and future projects include work on etch pipes, surface mounting and traceability of measurements.

This year's Quartz Devices Conference and Exhibition promises to be an important event for anyone involved with quartz technology. More information on the Quartz Devices Conference can be obtained by contacting the Electronic Industries Association, Components Group at 1722 Eye Street N.W., Washington, DC 20006. The telephone number is (202) 457-4930.

An Analysis of Inverter Crystal Oscillators

Improving Performance of a Common Circuit

By Leonard L. Kleinberg
NASA Gordon Space Flight Center

The ordinary linearly biased inverter is commonly employed in crystal oscillator circuits, such as the Pierce oscillator, up to several megahertz. However, by utilizing the inverter in a less conventional configuration, this frequency may be extended by a decade. The crystal may be either a fundamental or overtone cut.

The circuits are easily assembled since they have few components, are low cost, and exhibit better stability than the customary Pierce oscillator. The required supply is approximately 0.5 mA per megahertz.

The most commonly employed method of analyzing oscillator circuits is the feedback method. If an amplifier's gain is $k(\omega)$ and the feedback ratio is $B(\omega)$, the condition of oscillation is:

$$B(\omega)\,k(\omega) = 1 \tag{1}$$

For the circuit in Figure 1, which may be obtained by successive applications of the Thevenin/Norton theorem on a more common circuit,

$$B(\omega) = \frac{Z_2}{Z_1 + Z_2} = \frac{1}{k(\omega)} \tag{2}$$

Another method for analyzing oscillators is to look into the input terminal and equate the driving point admittance (DPA) to zero.

$$Y_{in} = DPA = \frac{1}{Z_2} + \frac{1 - k(\omega)}{Z_1} = 0 \tag{3}$$

$$DPA = \frac{1}{Z_2} + \frac{1}{Z_1} = \frac{k(\omega)}{Z_1} \tag{4}$$

Note that equation 4 is identical to equation 2. This second method of analysis is what will be used for the gate oscillators.

The circuit of Figure 2(a) is the most used gate crystal oscillator, where the crystal is operated in the inductive mode (L_x). Figure 2(b) depicts the approximate equivalent circuit used in determining the driving point admittance at the input terminal. The addition of R_1 is required, since the output resistance of the inverter is usually very small compared to the reactance of C_1. Direct application of equation 4 shows that there is no negative conductance in the DPA, and no oscillations can take place.

The DPA is given by:

$$DPA = \left[1 + \left(\frac{k}{R_1}\right)\left(\frac{1}{J\omega C_1}\right)\right]\left(\frac{1}{J\omega L_x + 1/(J\omega C_1)}\right) + \left(\frac{1+k}{R_B}\right)$$

$$= \frac{1}{J\omega L_x + 1/(J\omega C_1)} + \frac{k}{R_1(1 - \omega^2 L_x C_1)} + \frac{1 + k}{R_B} \tag{5}$$

To complete the DPA, C_2 is added in parallel to term 1 in equation 5:

(Condition 1)

$$J\omega C_2 + \frac{1}{J\omega L_x + 1/(J\omega C_1)} = 0 \tag{6}$$

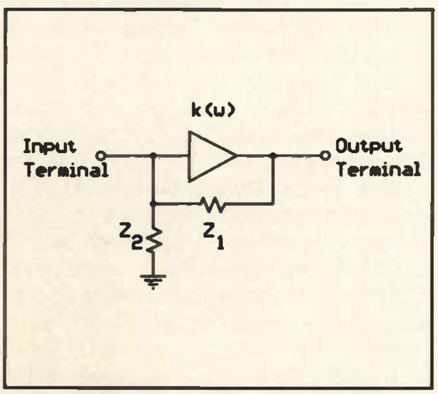

Figure 1. A simplified feedback circuit.

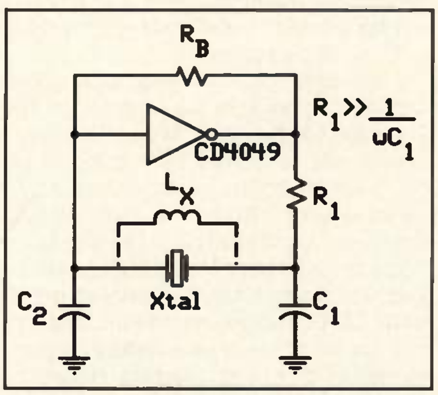

Figure 2(a). A commonly used gate crystal oscillator.

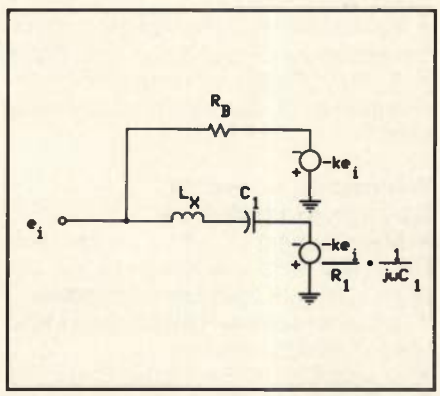

Figure 2(b). Approximate equivalent circuit to determine DPA.

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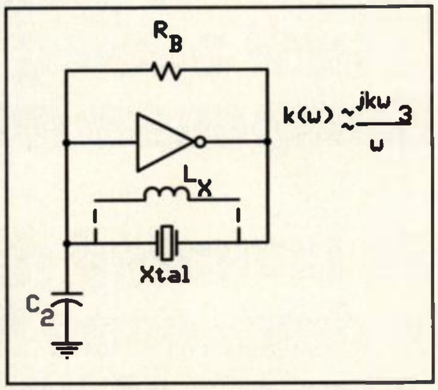


Figure 3(a). Circuit with 90 degrees phase shift.

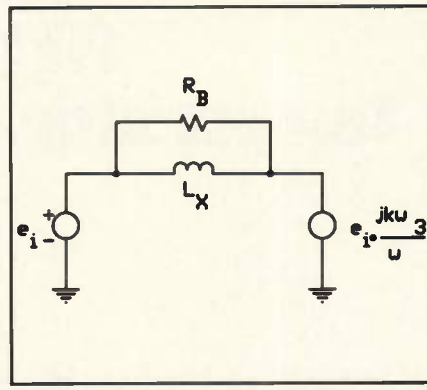


Figure 3(b). Equivalent circuit for Figure 3(a).

This yields the expected expression for ω^2:

$$\omega^2 = \frac{C_1 + C_2}{L_x C_1 C_2} \tag{7}$$

Substituting equation 7 into term 2 of equation 5 yields a negative conductance (G):

(Condition 2)

$$G = -\frac{kC_2}{R_1 C_1} = -\frac{1 + k}{R_B} \tag{8}$$

$$\frac{kC_2}{R_1 C_1} + \frac{1 + k}{R_B} = 0 \tag{8a}$$

It is interesting to note that if the circuit of Figure 2(a) is analyzed conventionally, a value of $B(\omega)$ equal to $-C_1/C_2$ is obtained. This implies that as C_2 is increased, the amount of feedback is reduced. However, equation 8 indicates the contrary — as C_2 is increased, the amount of feedback is increased. The latter case is the correct one.

The negative conductance, or positive feedback, was obtained because the voltage across C_1 experienced a 90 degree phase shift after the initial amplifier inversion. At low frequencies the R_1–C_1 elements are essential to oscillation, while at frequencies considerably above the 3 dB cut-off frequency R_1–C_1 may be omitted from the circuit, as the amplifier introduces the required 90 degree phase shift. The circuit is shown in Figure 3(a) and the equivalent circuit in Figure 3(b).

The DPA is given by:

$$\begin{aligned} DPA &= \left(1 - \frac{Jk\omega_3}{\omega}\right)\left(\frac{1}{J\omega L_X} + \frac{1}{R_B}\right) \\ &= \frac{1}{J\omega L_x} - \frac{k\omega_3}{\omega^2 L_x} + \frac{1}{R_B} \\ &\quad - \frac{Jk\omega_3}{\omega R_B} \end{aligned} \tag{9}$$

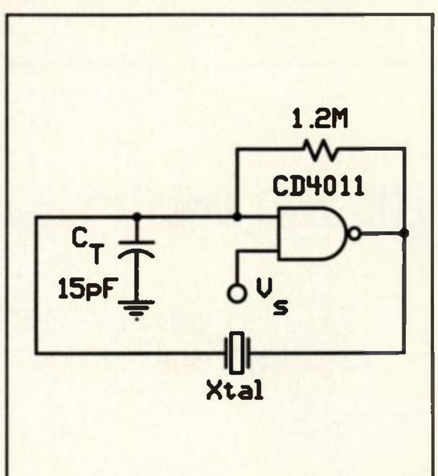


Figure 4. Basic circuit using the 4011.

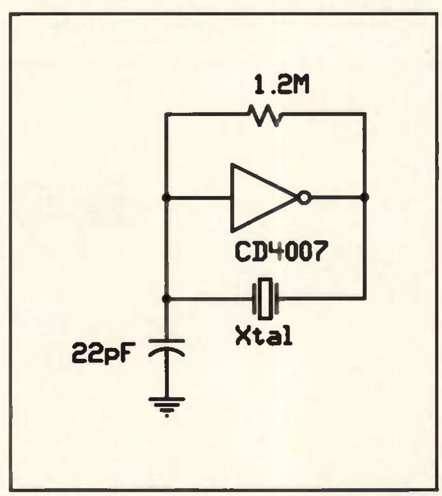


Figure 5. Basic circuit using the 4007.

For the case where R_B is very large (>1 Megohm), term 4 may be neglected noting that 4 is an inductive term and will be considered later. By adding C_2 at the input terminal, the design of the oscillator is completed.

$$J\omega C_2 + \frac{1}{J\omega L_x} = 0 \quad ; \quad \omega^2 = \frac{1}{L_x C_2} \qquad (10)$$

$$G = -\frac{k\omega_3}{\omega^2 L_x}$$

$$\frac{k\omega_3}{\omega^2 L_x} = \frac{1}{R_B} = k\omega_3 C_2 \qquad (11)$$

The gain vs. frequency response was plotted for the CD4007 and CD4049 (inverters) and for the CD4011 (2 input NAND), as a function of supply voltage. The values are shown in Tables 1, 2 and 3. The basic circuit using the CD4011 is depicted in Figure 4.

If an overtone crystal is used in the basic circuit (Figure 5), it will oscillate at the fundamental frequency. By adding a capacitor in series with the crystal, overtone operation is achieved. With today's newer crystals, in most cases a fundamental cut crystal may be obtained.

Another method of obtaining overtone operation, without C_1, is to parallel C_T with an inductor (L_1) such that:

$X_{L1} > X_{CT}$ at fundamental
$X_{L1} < X_{CT}$ at third overtone

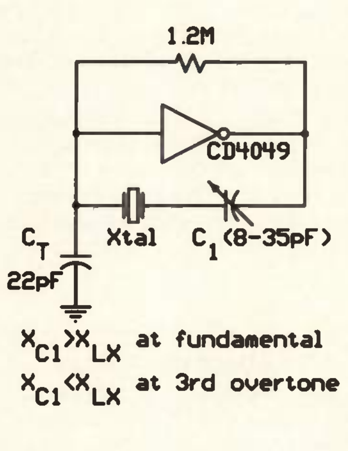


Figure 6. Circuit using the 4049.

V_{Supply}	k	F_3	Xtal Freq
5	50	300 kHz	5 MHz
10	25	2 MHz	10-20 MHz
15	20	2.2 MHz	20-50 MHz

Table 1. Gain vs. frequency response for the CD4049.

V_{Supply}	k	F_3	Xtal Freq
5	42	100 kHz	0.5-1 MHz
10	35	100 kHz	2 MHz
15	24	600 kHz	5 MHz

Table 2. Gain vs. frequency response for the CD4007.

V_{Supply}	k	F_3	Xtal Freq
5	20	800 kHz	5 MHz
10	20	2.0 MHz	10 MHz
15	20	2.2 MHz	15 MHz

Table 3. Gain vs. frequency response for the CD4011.

Using this condition of adding L_1, and referring to the circuit of Figure 3(a), the analysis of the DPA (equation 9) indicates that the fourth term $(-Jk\,\omega_3)/(\omega R_B)$ is an inductive admittance. By selecting R_B properly, it may serve the purpose of L_1 in Figure 7. The conditions are:

$$\frac{\omega_1 R_B}{\omega_3 k} < \frac{1}{\omega_1 C_T} \qquad \omega_1 \text{ is the fundamental crystal} \qquad (12)$$

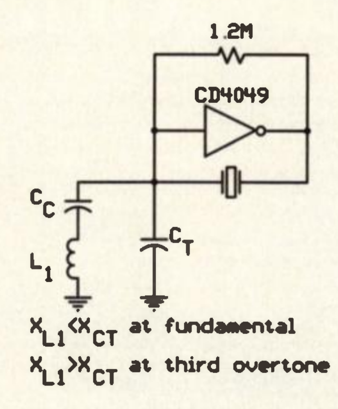


Figure 7. This circuit obtains overtone operation with the use of a parallel inductor.

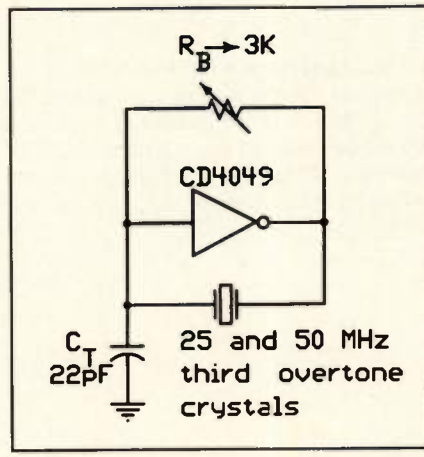


Figure 8. A simple third overtone oscillator.

$$\frac{3\omega_1 R_B}{\omega_3 k} > \frac{1}{3\omega_1 C_T} \qquad (13)$$

The circuit shown in Figure 8 oscillated at the third overtone of a 25 MHz and a 50 MHz crystal. Notice the simplicity of the circuit.

Summary

By biasing a CMOS inverter in its linear region and by using the device considerably beyond its 3 dB frequency, a simple, inexpensive and stable crystal oscillator — fundamental or overtone — may be quickly designed and assembled. The analysis describes the conditions of oscillation (i.e., the DPA is zero), and considers the gain of the inverter at high frequencies to be a 90 degree phase shifter. The technique is not limited to gate inverters, but includes op amps, transistors and FETs.

For a reasonably good time-base generator, the CD4049 with a 5 V supply and 2 MHz crystal (Figure 5) is approximately one part in 10^8 (1 sec). The 5 V supply should be derived by using a zener diode or a three terminal regulator. This regulation is good practice for this class of oscillators. For higher frequency oscillators the author finds it easier to work with fundamental cut crystal. rf

About the Author
Len Kleinberg is an electronic engineer at the NASA Gordon Space Flight Center, Code 728, Greenbelt, MD 20771. He can be reached by telephone at (301) 286-5683.

A Quartz Watch Time Base Monitor

By George P. Vella-Coleiro
AT&T Bell Laboratories

The circuit described here was developed to enable the frequency of oscillation of a quartz watch time base to be measured accurately in a non-invasive manner. Most digital watches on the market today use a quartz oscillator with a frequency of 32.768 kHz (2^{15} Hz) which is capable of great precision. The less expensive watch models, however, are typically adjusted at the factory to only ±30 s/month. Using the circuit described here and a simple frequency counter, it is possible to adjust a quartz watch to an accuracy of ±1 s/year.

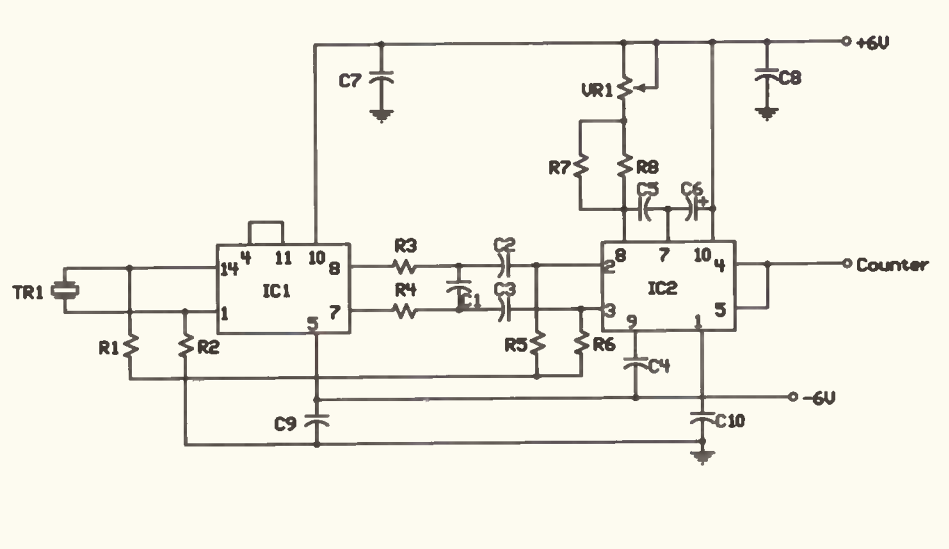
The problem with monitoring the frequency of a quartz watch arises from the very low power nature of the circuit. This means that even a high-impedance probe produces an excessive amount of loading. This problem is circumvented by using the acoustical energy emitted by the crystal. As might be expected, the acoustical power is minute, resulting in two difficulties that need to be overcome: interference and noise.

In the circuit shown in Figure 1, the problem of interference is taken care of by using a differential video amplifier (IC1) with good common mode rejection to amplify the weak signal picked up by the ultrasonic microphone (a piezoelectric transducer). This amplifier has a voltage gain of approximately 400, which is adequate to drive the input of phase-locked loop (PLL) IC2. The PLL takes care of the second problem, noise, by means of a loop filter with a long time constant. This effectively produces a very narrow noise bandwidth.

A PLL was chosen for this application instead of a narrow-band filter tuned to 32.768 kHz in order to avoid the problem of drift of the center frequency of the filter. It is estimated that achieving an adequate signal-to-noise ratio would require a filter with a bandwidth less than 1 Hz, necessitating a center frequency stability on the order of one part in 10^5. This level of stability is achievable through temperature compensation of the components that determine the center frequency, together with temperature control, but a PLL provides a much simpler solution. The frequency of oscillation of the voltage-controlled oscillator (VCO) in a well-designed PLL is locked to the frequency of the received signal, and the only requirement is that this frequency be within the capture range of the PLL.

A simple second-order loop filter, formed by C6 and a 3.6 kohm resistor internal to IC2, has been found to be entirely satisfactory for this application. The time constant was made large (about 0.1 s) so as to enhance immunity to noise and out-of-band signals. This was done at the expense of lock-up time and capture range, which are relatively unimportant in this application.

In Figure 1, video amplifier IC1 is configured for its maximum gain of 400, resulting in a bandwidth of 40 MHz. This bandwidth is limited to a much narrower range, centered around 30 kHz, by an R-C filter consisting of R3-R6 and C1-C3 before being applied to the input of PLL IC2. The simulated frequency response of the filter, including the output resistance of IC1 and the input resistance of IC2, is shown in Figure 2, indicating a 3 dB bandwidth of 10 kHz to 90 kHz. Figure 3(a) shows the watch signal



Parts List

R1	1 kohm
R2	1 kohm
R3	47 ohms
R4	47 ohms
R5	680 ohms
R6	680 ohms
R7	100 kohms
R8	3.9 kohms
VR1	200 ohms potentiometer
C1	0.022 µF
C2	0.018 µF
C3	0.018 µF
C4	0.0022 µF
C5	0.001 µF
C6	22 µF/10 V tantalum
C7-C10	0.1 µF ceramic
TR1	piezoelectric transducer (Radio Shack 273-064)
IC1	LM733CN video amplifier
IC2	LM565CN phase-locked loop

Figure 1. Circuit diagram of the quartz watch time base monitor.

obtained at the input of IC2 (pins 2 and 3) with IC2 removed from the circuit to avoid any pickup of the VCO output. The waveform at the output of IC2 (pin 4) with the PLL locked to the watch frequency is shown in Figure 3(b).

The free-running frequency of the VCO is adjusted by means of VR1 to within a few hundred hertz of 32.768 kHz in order to bring the frequency of the quartz oscillator within the capture range of the PLL. The watch is then placed atop the transducer, and the output of the VCO, monitored at pin 4 of IC2, becomes locked to the frequency of the quartz oscillator. The period rather than the frequency of the VCO is monitored by the counter in order to reduce the measurement time required to achieve the desired resolution. In a typical measurement, the counter is set to average 10^5 periods (the largest number available on the author's counter) and for a display where the least significant digit is 1 ps. This corresponds to a timing deviation of approximately 1 s/year. Successive readings of the counter fluctuate by several picoseconds, but averaging a number of readings reduces the fluctuations. A set of data where successive readings were averaged is shown in Figure 4. The abscissa shows the number of readings averaged, and the ordinate shows the average period expressed as a difference from the ideal value of 30,517,578 ps. A deviation of 1 ps corresponds to a timing error of approximately 1 s/year. Averaging 10 readings is sufficient to yield a measurement stable to better than ±1 ps, enabling the watch time base to be set to within 1 s/year.

The watch for which the data in Figure 4 were taken, an inexpensive model, gained about 4 s/week before adjustment. Since the frequency counter's time base was not calibrated to the required degree of accuracy (three parts in 10^8), the watch's drift rate was determined by comparison with standard time signals over a period of a few weeks. The requisite fractional change in period was then made using the frequency counter readings as relative rather than absolute measurements. After adjustment, the watch has gained less than 2 seconds in 3 months. Not bad for a timepiece costing less than $20!

The fact that a miniaturized, low power, inexpensive watch circuit can achieve such a high degree of accuracy is a tribute to the sophistication of modern integrated circuit technology. Undoubtedly, the frequency will drift over a period of several months, but the circuit described above is simple enough to use that adjustments can be made as needed to maintain the desired precision.

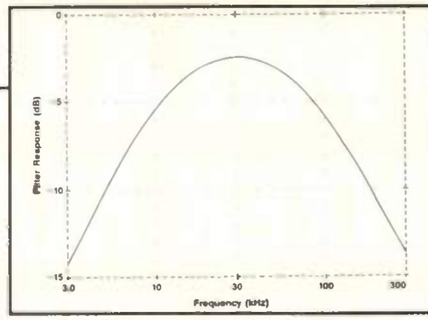


Figure 2. Simulated frequency response of the filter.

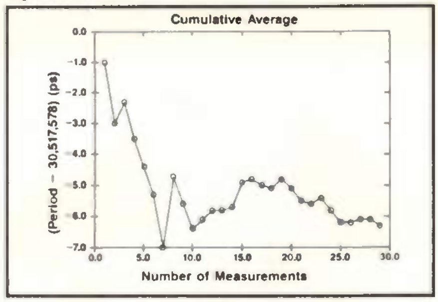


Figure 4. Period of oscillation of a quartz watch time base after adjustment.

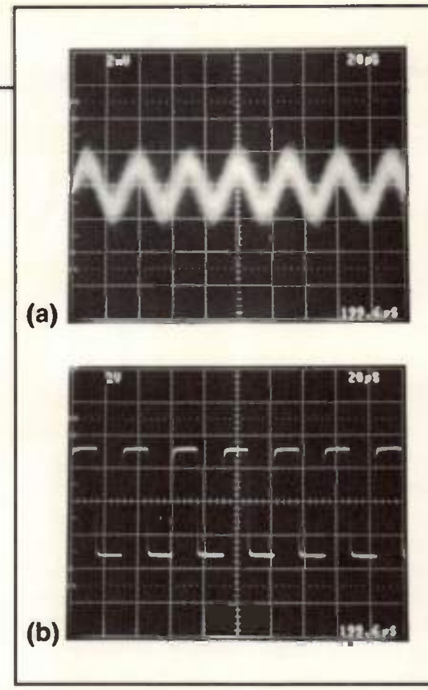


Figure 3. Waveforms observed at (a) the input of IC2(pins 2 and 3); and (b) the output of IC2 (pin 4).

About the Author

George P. Vella-Coleiro is with AT&T Bell Laboratories in Murray Hill, NJ 07974, where he supervises development of InP-based optoelectronic integrated circuits for fiber communications. His time base monitor design won one of the runner-up prizes in the 1989 RF Design Awards Contest. Mr. Vella-Coleiro can be reached at (201) 582-3381.

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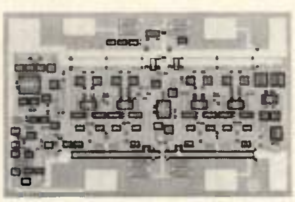

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Quartz Resonator Model Measurement and Sensitivity Study

By Donald C. Malocha, Huat Ng and Michael Fletcher
University of Central Florida

The article presented here was named "Best Paper" of the 10th Quartz Devices Conference and Exhibition, August 30 through September 1, 1988. The work described was supported by a grant from the Electronic Industries Association, Washington, D.C.

The results of a study of measurement techniques, model parameter sensitivity, and analysis of quartz resonator devices are presented in this article. A data reduction technique for one-port and two-port scattering parameter device measurements is implemented, along with computer analysis based on the quartz crystal model. The data reduction technique presented is significantly more accurate than the currently available measurement technique. Experimental and theoretical data for several quartz resonator device samples are shown, and results for several quartz crystal devices are presented using the analysis approach discussed.

The analysis presented here for one- and two-port quartz resonators is based on the EIA-512 data reduction technique. The technique was implemented using FORTRAN code on an IBM PC to extract the model parameters.

General Analysis for the One-Port Model

The simple equivalent circuit model used for the analysis of the one-port model is shown in Figure 1. The following analysis will use the admittance (Y) parameters. From the circuit, the input admittance is given by:

$$Y = jB_0 + \frac{R_1 - jX_1}{R_1^2 + X_1^2} \tag{1}$$

The conductance can be written as:

$$G = \frac{R_1}{R_1^2 + X_1^2} \tag{2}$$

and the susceptance can be written as:

$$B = B_0 - \frac{X_1}{R_1^2 + X_1^2} \tag{3}$$

Squaring equations 2 and 3 and rewriting yields:

$$G^2 + (B - B_0)^2 = 1/(R_1^2 + X_1^2) \tag{4}$$

Substituting R_1/G from equation 2 for the denominator of equation 4 yields:

$$G^2 + (B - B_0)^2 = G/R_1 \tag{5}$$

Completing the square and rewriting equation 5:

$$[G-1/(2R_1)]^2 + (B-B_0)^2 = [1/(2R_1)]^2 \tag{6}$$

Substituting $-(B-B_0)/X_1$ from equation 3 into equation 4, completing the square, and manipulating yields:

$$G^2 + [B-B_0 +1/(2X_1)]^2 = [1/(2X_1)]^2 \tag{7}$$

Both equation 6 and equation 7 are in the form of an equation of a circle. The motional arm parameters, R_1 and X_1, are separated. In order to determine the resonator parameters, these equations must be solved simultaneously at a given frequency. This solution corresponds to the point or points where the two circles intercept at a particular frequency. Rather than attempt to solve at any one frequency, the analysis proceeds by making many measurement points and then determining the best fit to the parameters statistically.

Although equations 6 and 7 are in the form of an equation of a circle, these will not plot a circle unless G_0, B_0 and X_1 are constant with frequency. G_0 represents a static conductance and is frequency-independent based on the resonator model.

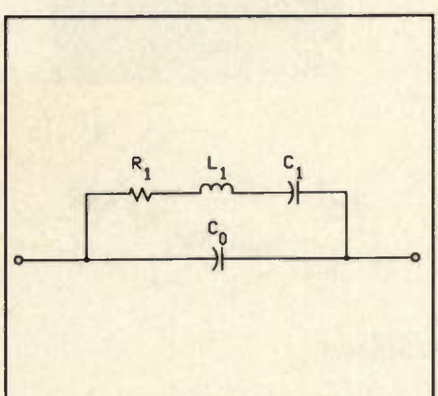

Figure 1. Single-mode one-port resonator equivalent circuit model.

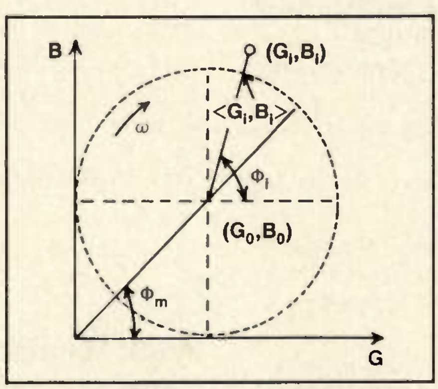

Figure 2. Conductance-susceptance plane for an ideal quartz resonator.

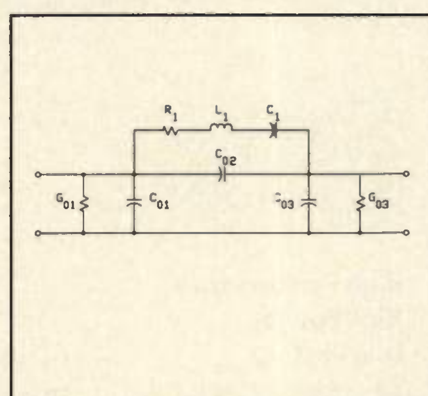

Figure 3. Single-mode two-port resonator equivalent circuit model.

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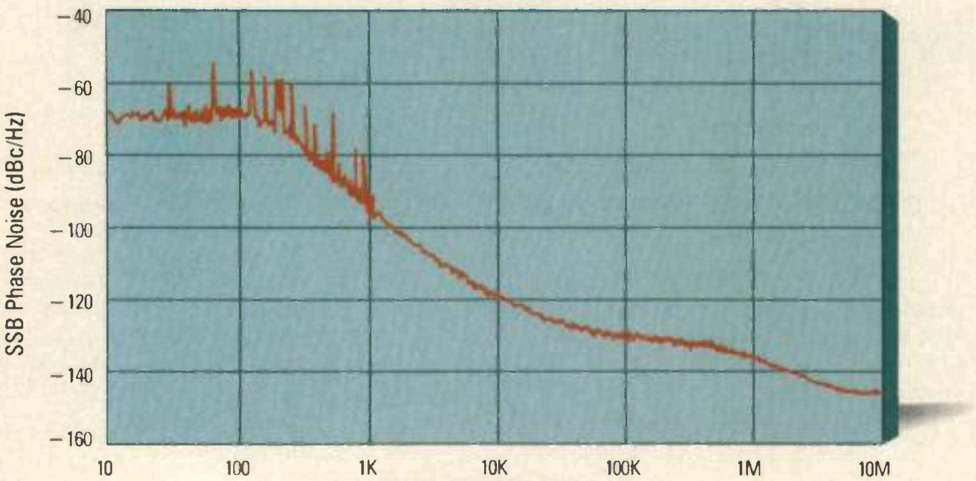

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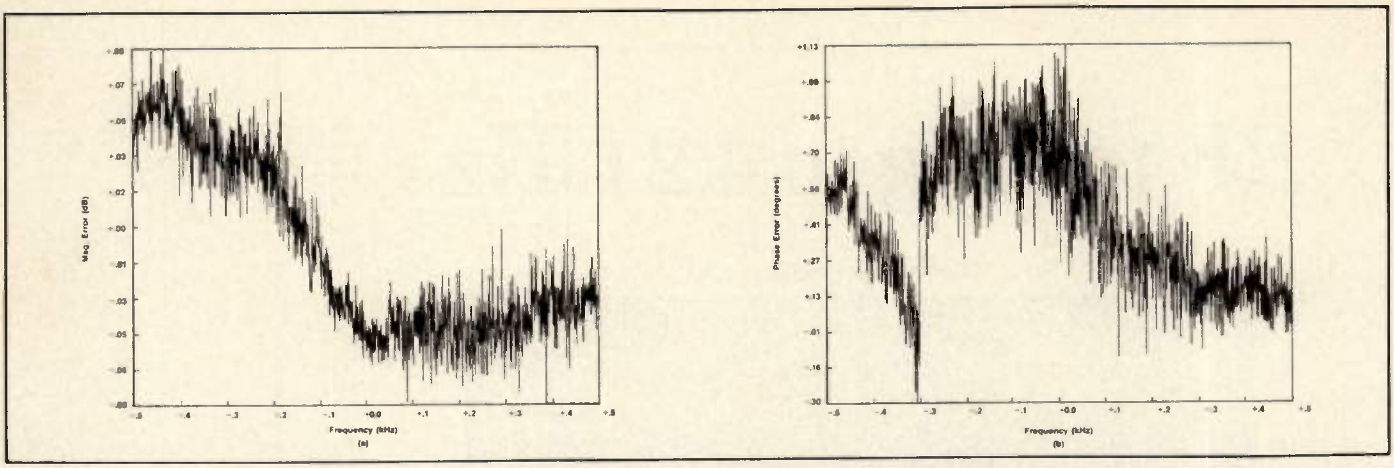

Figure 4. Amplitude (a) and phase (b) of one-port S_{11} resulting from the difference of two consecutive runs.

X_1 is the motional arm reactance and is certainly frequency-dependent. B_0 is the static capacitance contribution to the susceptance, and is also frequency-dependent. However, it is desired to force equation 6 into the proper form for a circle. This is accomplished by making the narrowband approximation as follows:

$$B_0 = \omega C_0 \text{ and } B_s = \omega_s C_0$$

or,

$$B_0 = B_s \omega/\omega_s \quad \text{where } \omega = \omega_s \pm \delta\omega$$

Using the narrow bandwidth approximation, where $\delta\omega$ is small and considered negligible with respect to ω_s, B_0 is approximately equal to B_s, which is a constant with frequency. This is the first approximation in the analysis approach which will yield some error. Equation 6 is now in the proper form of a circle where the center is (G_0, B_0) and G_0 and B_0 are constants when plotted on the admittance plane. Using equation 6 and making the substitution $G_0 = 1/(2R_1)$ gives:

$$(G-G_0)^2 + (B-B_0)^2 = G_0^2 \qquad (8)$$

Expanding and then rearranging equation 8 yields:

$$G^2 + B^2 = 2G_0G + 2B_0B - B_0^2 \qquad (9)$$

Statistical Analysis

If exact measurements were possible, it would be unnecessary to perform a statistical analysis; however, this is not possible. Now, it is desirable to make several measurements of G and B at various frequencies around resonance. Let G_i and B_i be the conductance and susceptance measurements at the ith frequency. Next, a set of variables is defined as follows:

$$M_i = G_i^2 + B_i^2$$
$$K_1 = -B_0^2$$
$$K_2 = 2G_0$$
$$K_3 = 2B_0$$

Then, from equation 9,

$$M_i = K_1 + K_2G_i + K_3B_i \qquad (10)$$

where each K_n is a constant, independent of frequency. An error function can be written as:

$$\varepsilon = \sum_{i=1}^{I} (M_i - K_1 - K_2G_i - K_3B_i)^2 \qquad (11)$$

which represents the mean squared error from a set of measured data containing a total of I points. There will be some statistical error in the measured data points; it is therefore necessary to find the *best* coefficients, in order to minimize the error over all measured points. This is accomplished by finding the minimum error in ε with respect to K_n, given by:

$$\frac{\delta\varepsilon}{\delta K_1} = \sum_{i=1}^{I} 2\{Y_i - K_1 - K_2G_i - K_3B_i\} = 0 \qquad (12a)$$

$$\frac{\delta\varepsilon}{\delta K_2} = \sum_{i=1}^{I} 2\{Y_i - K_1 - K_2G_i - K_3B_i\}(-G_i) = 0 \qquad (12b)$$

$$\frac{\delta\varepsilon}{\delta K_3} = \sum_{i=1}^{I} 2\{Y_i - K_1 - K_2G_i - K_3B_i\}(-B_i) = 0 \qquad (12c)$$

Since these terms are linear in their coefficients, equation 12 can be re-written in matrix form. The matrix equation can be solved for the K_n coefficients using gaussian elimination and is easily programmed. Several of the model parameters are now determined from the coefficients, and are given by:

$$1/R_1 = K_2 \text{ or } G_0 = K_2/2$$
$$B_0 = K_3/2 \text{ or } B_0 = [-K_1]^{1/2} \qquad (13)$$

These are the first model parameters to be found and the approximation is that B_0 is constant over the frequency range of interest.

Expected Values of Conductance and Susceptance

Given the solutions for G_0 and B_0, it is now possible to find the expected values for every conductance and susceptance point as a function of frequency. Figure 2 shows a typical admittance plane circle. Hafner (1) discusses many of the important points of interest regarding the admittance plane circle as related to the crystal resonator. Several parameters are necessary to continue the analysis. First, an angle Θ_m is defined as:

$$\Theta_m = \tan^{-1}[B_0/G_0] \qquad (14)$$

where Θ_m represents the angle of the vector from (0,0) to (G_0, B_0), and can be used to find the maximum admittance or minimum impedance. Referring to Figure 2, an angle Θ_i is defined as:

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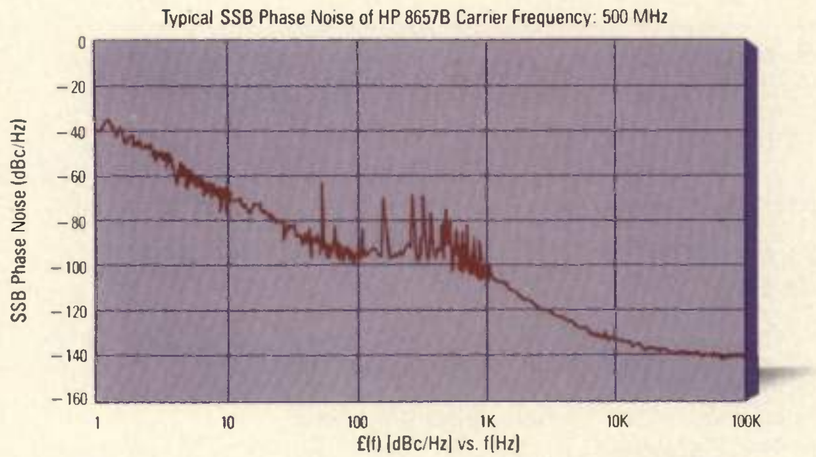

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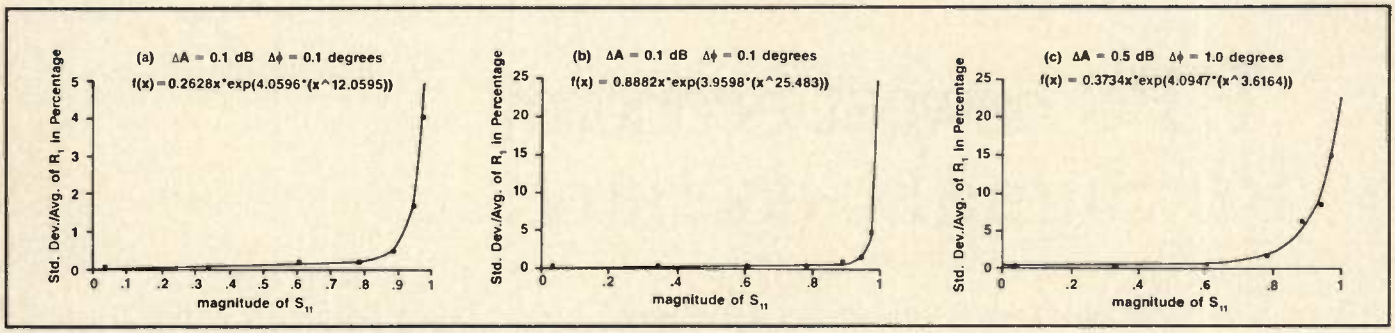

Figure 5. Simulations of parameter R_1 variations vs. $|S_{11}|$.

$$\Theta_i = \tan^{-1}\left[\frac{B_i - B_0}{G_i - G_0}\right] \tag{15}$$

where Θ_i represents the angle of the vector to the actual measured values of G_i and B_i.

The expected value of the conductance, $<G_i>$, and of the susceptance, $<B_i>$, at a given frequency, are given by:

$$\begin{aligned} <G_i> &= G_0 + \text{Rad} * \cos(\Theta_i) \\ <B_i> &= B_0 + \text{Rad} * \sin(\Theta_i) \end{aligned} \tag{16}$$

Now, the conductance and susceptance error, given as the difference of the expected value minus the measured value, is written as:

$$\begin{aligned} \gamma_{G_i} &= <G_i> - G_i = G_0 + \text{Rad} * \cos(\Theta_i) - G_i \\ \gamma_{B_i} &= <B_i> - B_i = B_0 + \text{Rad} * \sin(\Theta_i) - B_i \end{aligned} \tag{17}$$

Next, the error terms are squared, yielding:

$$\begin{aligned} \gamma_{G_i}^2 &= (<G_i>-G_i)^2 \\ &= (G_0-G_i)^2 + 2*\text{Rad}*(G_0-G_i)*\cos(\Theta_i) + \text{Rad}^2*\cos^2(\Theta_i) \end{aligned} \tag{18a}$$

$$\begin{aligned} \gamma_{B_i}^2 &= (<B_i>-B_i)^2 \\ &= (B_0-B_i)^2 + 2*\text{Rad}*(B_0-B_i)*\sin(\Theta_i) + \text{Rad}^2*\sin^2(\Theta_i) \end{aligned} \tag{18b}$$

Equation 18 represents the squared error between the expected values based on a best fit to the circle and the actual measured value at a given frequency. The statistical variance can now be found from these expressions, and is given as:

$$\sigma_G^2 = \frac{\sum_{i=1}^{N} (\gamma_{G_i}/<G_i>)^2}{N} \tag{19a}$$

$$\sigma_B^2 = \frac{\sum_{i=1}^{N} (\gamma_{B_i}/<B_i>)^2}{N} \tag{19b}$$

A combined variance can also be defined as:

$$\sigma_{GB}^2 = \sigma_G^2 + \sigma_B^2 \tag{20}$$

where this would give a measure of the error sum.

Another approach is to measure the error of the expected value of the circle radius from that obtained based on a single measurement point. This is given as:

$$\gamma_R = \text{Rad}\sqrt{(G_i - G_0)^2 + (B_i - B_0)^2} \tag{21}$$

The variance of the radius of the ensemble is given by:

$$\sigma_R^2 = \frac{\sum_{i=1}^{N} (\gamma_R/\text{Rad})^2}{N} \tag{22}$$

This is the variance used to determine points which are in error due to bad measurement.

Determining L_1 and C_1

The last model parameters to be determined are the motional arm reactances, L_1 and C_1. Because X_1 is a nonlinear function of frequency, the parameters L_1 and C_1 are not as easily obtained as G_0. The approach is to approximate the frequency-dependent reactance, $X_1(f)$, using a series, and then determine the series coefficients using the least square approximation. The motional arm reactance at a given frequency, f_i, is obtained using equations 3 and 4, and is given by:

$$X_{1_i} = \frac{-<\bar{B}_i>}{<G_i>^2 + <\bar{B}_1>^2} \tag{23}$$

where $<\bar{B}_i> = <B_i> - B_0$ and $<G_i>$ and $<\bar{B}_i>$ are the expected values as derived previously. Now $X_1(f)$ can be fit to an nth-order power series:

$$X_1(f) = A_0 + A_1 f + A_2 f^2 + A_3 f^3 + \ldots \tag{24}$$

The higher the order of the equation, the better the fit of the equation to the values of $X_1(f)$ over the frequencies of interest. Normally, only a quadratic equation is used since this yields a good fit to $X_1(f)$ over narrowband frequencies around f_s. The coefficients, A_n, are obtained in a manner similar to that in which the K_n coefficients were found. An error function is defined as:

$$\varepsilon = \sum_{i=1}^{I} (X_i - A_0 - A_1 f_i - A_2 f_i^2)^2 \tag{25}$$

which represents the mean squared error from a set of measured data containing I points. The minimum error in ε with respect to A_n is given by:

$$\frac{\delta\varepsilon}{\delta A_0} = \sum_{i=1}^{I} 2\{X_{1_i} - A_0 - A_1 f_i - A_2 f_i^2\} = 0 \tag{26a}$$

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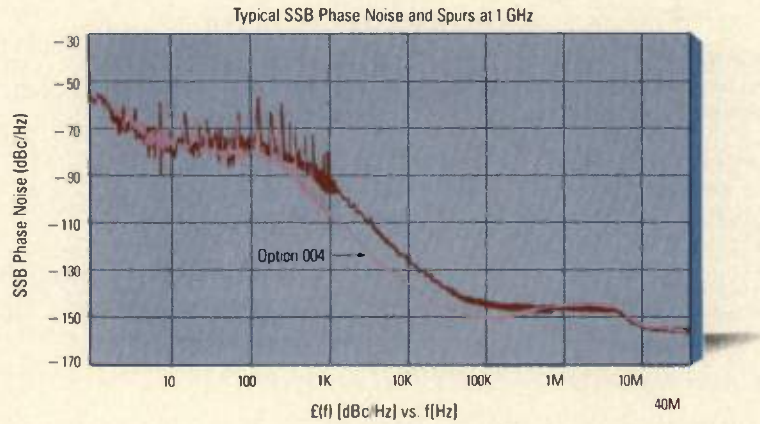

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$$\frac{\delta\varepsilon}{\delta A_1} = \sum_{i=1}^{I} 2\,\{X_{1_i} - A_0 - A_1 f_i - A_2 f_i^2\}\,(-f_i) = 0 \tag{26b}$$

$$\frac{\delta\varepsilon}{\delta A_2} = \sum_{i=1}^{I} 2\,\{X_{1_i} - A_0 - A_1 f_i - A_2 f_i^2\}\,(-f_i^2) = 0 \tag{26c}$$

Since these terms are linear in their coefficients, equation 24 can be re-written in matrix form. The matrix equation can be solved for the A_n coefficients using gaussian elimination and is easily programmed.

At series resonance, $X_1(f_s) = 0$ and the susceptance is equal to B_0. Having solved for the A_n coefficients, equation 24 can be used to determine X_1 at any frequency. Therefore, to find f_s, the following quadratic equation is solved:

$$f_s = \frac{-A_1 + (A_1^2 - 4A_2A_0)^{1/2}}{2A_2} \tag{27}$$

Now that f_s is known, the motional arm reactances can be found. The reactance X_1 is given by:

$$X_1 = \omega L_1 - 1/(\omega C_1) = (\omega^2 L_1 C_1 - 1)/(\omega C_1) \tag{28}$$

At resonance, $f = f_s$ and $X_1 = 0$, which yields, from equation 28,

$$\omega_s = (L_1 C_1)^{-1/2} \tag{29}$$

Next, take the derivative of X_1 with respect to ω and evaluate at ω_s:

$$\left.\frac{dX_1}{d\omega}\right|_{\omega=\omega_s} = \frac{2(\omega_s L_1 C_1)\omega_s C_1 - C_1(\omega_s^2 L_1 C_1 - 1)}{(\omega_s C_1)^2} \tag{30}$$

Substituting equation 29 into equation 28, substituting $\omega = 2\pi f$, and also using equation 24 yields:

$$L_1 = \frac{1}{4\pi} \bullet \left.\frac{dX_1}{df}\right|_{f=f_s} = A_1 + A_2 f_s \tag{31}$$

The motional capacitance can now be determined as:

$$C_1 = 1/(4\pi^2 f_s^2 L_1) \tag{32}$$

Finally, C_0 is simply given by:

$$C_0 = B_0/(2\pi f_s) \tag{33}$$

At this point all of the resonator crystal parameters are found.

Discussion

The ideal quartz crystal resonator model analysis and data

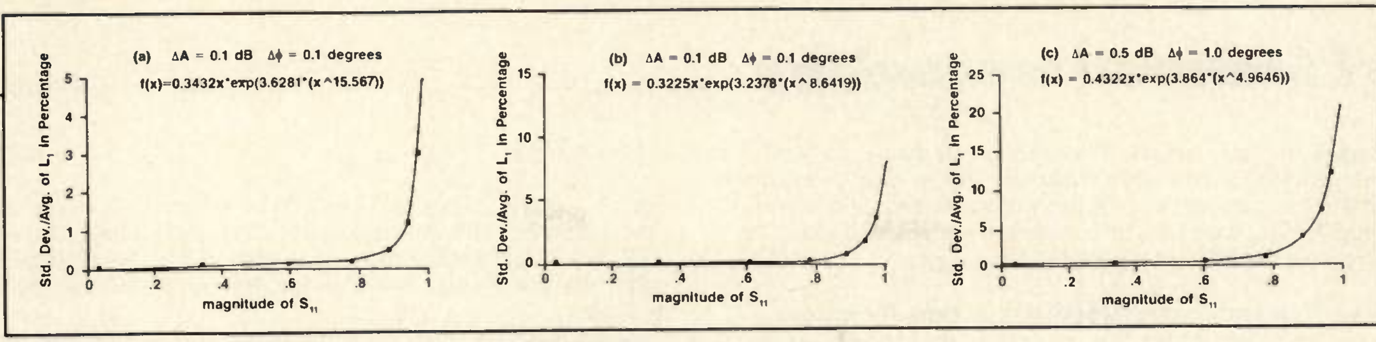

Figure 6. Simulations of parameter L_1 variations vs. $|S_{11}|$.

reduction technique are quite good. The statistical approach for determining the model parameters is accurate, and helps to remove individual measurement errors, if properly implemented.

There are only two approximations used in the statistical analysis which can give rise to errors. The first approximation assumes that the susceptance due to the static capacitance is constant with frequency. Its value is set equal to $\omega_0 C_0$, which is constant. For high Q resonators, this is a small perturbation over frequency and its effect gives rise to extremely small errors in the model data. This error can be reduced greatly by determining $\omega_0 C_0$, as described in the analysis, in a first run and then compensating for its frequency dependence in a second run. It is straightforward to compensate for the frequency-varying susceptance in the analysis; however, there is a slight disadvantage due to the longer computational run time. This iterative technique was tested and significant change in the extracted model parameters was not noted.

The second approximation occurs in fitting the motional arm reactance to a second-order polynomial. From the results obtained in our analysis, the resonant frequency is found quite accurately; however, the motional arm parameters seem to have a greater variance than anticipated. These results are seen in the data presented. The motional arm inductance is found by taking the derivative of the reactance with respect to frequency, evaluated at series resonance. It seems reasonable that the inductance would be more sensitive to the order of the polynomial, since it is varying with respect to the slope at a given point. This sensitivity would probably be reduced by expanding the polynomial series to the third order. This option has not been implemented in the present computer code or models.

There are also unavoidable round-off errors in any calculation due to the number of significant digits used. In the computer program implementation, nearly every variable is in double precision to maintain the greatest number of significant digits in the calculations. The areas of the analysis which have the greatest potential for error are in the matrix equations where the polynomial equation constants are determined. In order to minimize these errors, several variables and the matrix elements are scaled for greatest accuracy.

Two-Port Resonator Model Extraction Technique

In order to determine or isolate the quartz resonator parameters from the package electrical parasitics, the one-port model is extended to a two-port equivalent circuit model, as shown in Figure 3. The external capacitances and conductances are principally due to the package pin out configuration and structure. They may also include fixturing effects due to insertion of the device, which results in contact resistance and fringing capacitances.

The network equation can be written in matrix form as:

$$\begin{vmatrix} I_1 \\ \\ I_2 \end{vmatrix} = \begin{vmatrix} G_{01}+j\omega C_{01} + Y_c + j\omega C_0 & -Y_c - j\omega C_0 \\ -Y_c - j\omega C_0 & G_{02}+j\omega C_{02} + Y_c + j\omega C_0 \end{vmatrix} \begin{vmatrix} V_1 \\ \\ V_2 \end{vmatrix} \quad (34)$$

where Y_c represents the quartz resonator motional terms.

The admittances are defined as:

$$Y_{11} = G_{01} + j\omega C_{01} + Y_c + j\omega C_0$$
$$Y_{12} = Y_{21} = -Y_c - j\omega C_0$$
$$Y_{22} = G_{02} + j\omega C_{02} + Y_c + j\omega C_0 \quad (35)$$

Both Y_{12} and Y_{21} have the ideal crystal resonator admittance model isolated from the package and parasitic terms. Therefore, the quartz crystal model parameters can be extracted using the exact same equation as derived for the one-port analysis.

Given exact Y_{12} and Y_{21} admittances, either term can be used to extract the model parameters. A better approach would be to average Y_{12} and Y_{21}. Assuming uncorrelated random errors, this will have the effect of minimizing random errors, thereby increasing accuracy.

The package and parasitic elements may now be determined from:

$$G_{01} + j\omega C_{01} = Y_{11} + Y_{21}$$
$$G_{02} + j\omega C_{02} = Y_{22} + Y_{21} \quad (36)$$

Given that there are N sampled points, the parameters can be found using the statistical average, given by:

$$G_{01} = \sum \mathrm{Re}\,\{Y_{11} + Y_{12}\}\,/N \quad (37a)$$

$$C_{01} = \sum \mathrm{Im}\,\{Y_{11} + Y_{12}\}\,/(\omega N) \quad (37b)$$

$$G_{02} = \sum \mathrm{Re}\,\{Y_{22} + Y_{21}\}\,/N \quad (37c)$$

$$C_{02} = \sum \mathrm{Im}\,\{Y_{22} + Y_{21}\}/(\omega N) \quad (37d)$$

All elements of the two-port model are determined. The calculations and computer programming require only minor modifications of the one-port model.

The advantage of the two-port analysis is the extraction of several parasitic parameters. This is necessary if a more exact model of the actual crystal parameters is required. It should, however, be stated under what conditions the model was measured and what model is being used. Otherwise, the parameters can be a source of confusion for the end-user.

Computer Programs and Data Acquisition

Computer programs were developed to implement the data extraction technique and to run sensitivity. The programs were written in FORTRAN on IBM PC compatible computers, using a math co-processor. All variables used double-precision (16 bit) format for greatest accuracy. The programs written are menu-driven and were structured to handle multiple inputs of files where necessary, to aid in compiling data and statistics. The programs written include: an ideal resonator admittance model to S-parameter conversion; one-port and two-port model extraction programs; and a random number program for perturbation of S-parameters.

In addition, several programs were written to obtain S-parameters vs. frequency data using an HP 3577 automatic network analyzer (ANA), an HP 9845 and an IBM PC compatible for the data acquisition. The one-port data acquisition used the four-term error correction technique and the two-port used the ten-term error correction technique (3). Crystal data was taken carefully; however, no environmental controls were maintained on the crystals for the data presented. The crystal test set included open, short, through and Z_0 calibrations and conforms to recommendations of EIA-512. The S-parameter data is converted to Y-parameters in each of the analysis programs.

Experimental Data

In order to understand the limitations and problems of a typical measurement system and test set, and to provide data for confirmation of the models developed, data was obtained on several quartz resonators. Several devices were tested over an extended period, while others were tested over a period of only a few minutes. No averaging was used in data acquisition.

Figure 4 shows a representative sample of the results of subtracting two consecutive, one-port S-parameter runs of a 40 MHz quartz crystal with bandwidth equal to 1 kHz. There appear to be two sources of error. The first appears to be random noise associated with noise in the measurement system. The second is a slowly varying variation, probably due to short-term environmental changes. There is also a discontinuity in phase which has been seen on various runs. These errors give rise to statistical variations in the extracted resonator parameters. The measurement system is probably representative of typical industrial systems.

Results were obtained for two quartz resonators, a 40 MHz resonator labeled F81 and a 3.6 MHz resonator labeled J69. The extracted one-port model parameters for F81 and J69 were calculated. The number of points beyond the $2.7\sigma_r^2$ variance when fitting the circle (equation 22) were noted. The data is fairly consistent despite time and small temperature variations. It is also consistent with results provided by the EIA "Round Robin" experiments (4).

A number of one-port measurement runs were conducted to determine the effects of the number of data points taken and the measurement bandwidth. Over reasonable sample points ($16 \leqslant NUM \leqslant 128$) and measurement bandwidth, no significant statistical variations were noted. Therefore, using 16 as the number of sample points and the crystal's approximate 3 dB bandwidth should give good results. This was also verified by using theoretically generated S-parameters from an ideal one-port quartz resonator model. Statistical variation in σ_r for the theoretical data was found to be less than 0.1 ppm, as compared to data for the extracted one-port model parameter measurements, which showed several thousand parts per million variations.

$R_1 = 25\ \Omega$, $L_1 = 9.82$ mH, $C_1 = 0.00156$ pF, $C_0 = 5$ pF, $f_s = 40.663250$ MHz

ΔA, $\Delta\phi$ (dB) (deg)	R_1 (10^{-4})	L_1 (10^{-4})	C_1 (10^{-4})	C_0 (10^{-4})	σ_r (10^{-6})
0.05, 0.1	2.5	0.95	0	19.5	1,027
0.1, 0.5	4	2.85	2.17	39.5	2,035
0.5, 0.5	24	7.84	7.7	192	10,364
0.5, 1.0	27	9.78	9.6	242	10,145
1.0, 1.0	49	10.4	10.7	366	20,965

Table 1. Summary results of random variations on one-port resonator.

Data was also taken using the two-port measurement techniques and data extraction models. Results were obtained for consecutive runs over a one-hour period with the network analyzer resolution bandwidth (RBW) as the variable. For F81, the model parameters are consistent. The variation in σ_r is less than 1000 ppm and is better than that obtained for the one-port model. The approximate bandwidth is 3 kHz and the effect of the RBW is small. Results for J69, the 3.6 MHz crystal, show both G_{01} and G_{03} are small and negative and can be assumed zero. The effect of RBW is evident when RBW = 1 kHz, which is greater than the filter bandwidth of approximately 200 Hz. Results indicated that as the RBW approaches the resonator bandwidth, errors in the extracted model parameters occur.

However, it should also be noted that the magnitude of the changes in the model parameters for this case are very small, and no change in center frequency is seen. Comparison of results obtained shows that the σ_r for the two-port measurements is better than that for the one-port by approximately a factor of 10. It appears that better results are obtainable with the two-port vs. the one-port measurement technique.

Model Parameter Sensitivity

Simulations of the effects of random errors were performed by perturbing theoretically obtained S-parameters by random variations. The S-parameters were generated by using the ideal one-port resonator equivalent circuit model. The random variations were obtained using a random number generator with a given variance. The theoretical file was modified 15 times using a different "seed" for the random number generator to generate the statistical variations. A greater number of runs (about 50) are necessary to yield better accuracy; however, the data obtained was very good since all averages (R,L,C) approached the mean theoretical value. The standard deviation of the admittance circle fit is used as a good figure of merit. For a standard deviation in amplitude of 0.1 dB and in phase of 0.5 degrees, the average deviation σ_r is approximately equal to 2×10^{-3}. This is in fairly good agreement with the data obtained for the F81 one-port model, and yields a measure of the magnitude of errors which may be seen in the measurement system. Also of note, the determination of center frequency was unchanged regardless of the magnitude of the random deviations on the S-parameters. A summary of results for this single device simulation is given in Table 1.

Finally, simulations were conducted on the variations of model parameters vs. the magnitude of S_{11} for varying random deviations on theoretically generated S-parameters. R_1 was varied to obtain changes in $|S_{11}|$ while all other parameters were kept constant. S_{11} was calculated and then random amplitude and phase deviations were added to the theoretically generated S-parameter data. A random number run was generated to produce the plots. Variations in R_1 and L_1 are shown in Figures 5 and 6. Although this was a simple approach, it can be concluded that for a fixed random variation in amplitude and phase measurements, the deviation in parameter values increase as $|S_{11}|$ increases.

(The experimental data described here is reproduced in full in the Proceedings of the 10th Quartz Devices Conference, 1988.)

Conclusion and Discussion

From the measurement data and theoretical analysis it is concluded that the data reduction technique presented is significantly more accurate than the currently available measurement technique. The center frequency can be obtained to better than 1 part per million, based on theoretical simulations. Greater accuracy may be obtainable when using 32 bit computer systems, but this was not investigated. Both one-port and two-port resonator model parameter extraction approaches were investigated and both were determined to be accurate. Measurements on samples of quartz resonators were presented which indicate the range of parameter deviations. The deviations arise from several possible phenomenon: noise inherent in the measurement system; non-ideal calibration; environmental fluctuations; and random discontinuity due to system contacts, settling time, etc. Although these can or do all exist in a measurement, by removing bad points and using the statistical analysis approach the various crystal parameters can be extracted with a high degree of accuracy. Based on the test set used, accurate reproducible measurements can be taken. Experimental measurements indicate that this test set-up can obtain measurement accuracy to approximately ±0.1 dB and ±0.8 degrees. This does not assure that this consistency of parameters or accuracy in data acquisition can be found from site to site.

Using the theoretical approach, it appears that parameter variations are correlated to the magnitude of S_{11} and the random measurement deviations for a one-port device. This has not been extended to the two-port analysis; however, it seems reasonable that similar results will be found.

The two-port model has the advantage of isolating some of the parasitic elements from the ideal crystal model, as compared to the one-port model. In addition, time variant fixture effects, most notably contact resistance and fringing capacitance, should be isolated in the parasitic model elements rather than in variations of the crystal parameters.

Although this article has not discussed the measurement calibrations, this is worth some discussion. For the one-port model, only four error terms are generated and the equations with terminations are relatively manageable. Therefore, correction for frequency-dependent terminations is possible. For the two-port, ten-term model, it is very impractical to attempt to correct for frequency-dependent terminations. Therefore, the data measurements are only as good as the terminations used. This could lead to device measurement variations from site to site.

Acknowledgements

The authors would like to thank Rami Rifai who obtained some of the quartz resonator device measurements; Sallie Douglas, who wrote portions of the data acquisition routines; and Jeanne Gallagher, who greatly aided in the drafting of this paper. The authors would also like to express appreciation to Canon Bradley, Warren Smith, Jack Kusters, and Jack Kay for many fruitful discussions and guidance in this work, and to Warren Smith for providing the test fixture and sharing of data and reports from his testing at various sites. Quartz resonator devices were provided by Mr. Warren Smith, consultant, and by Colorado Crystal Corporation. rf

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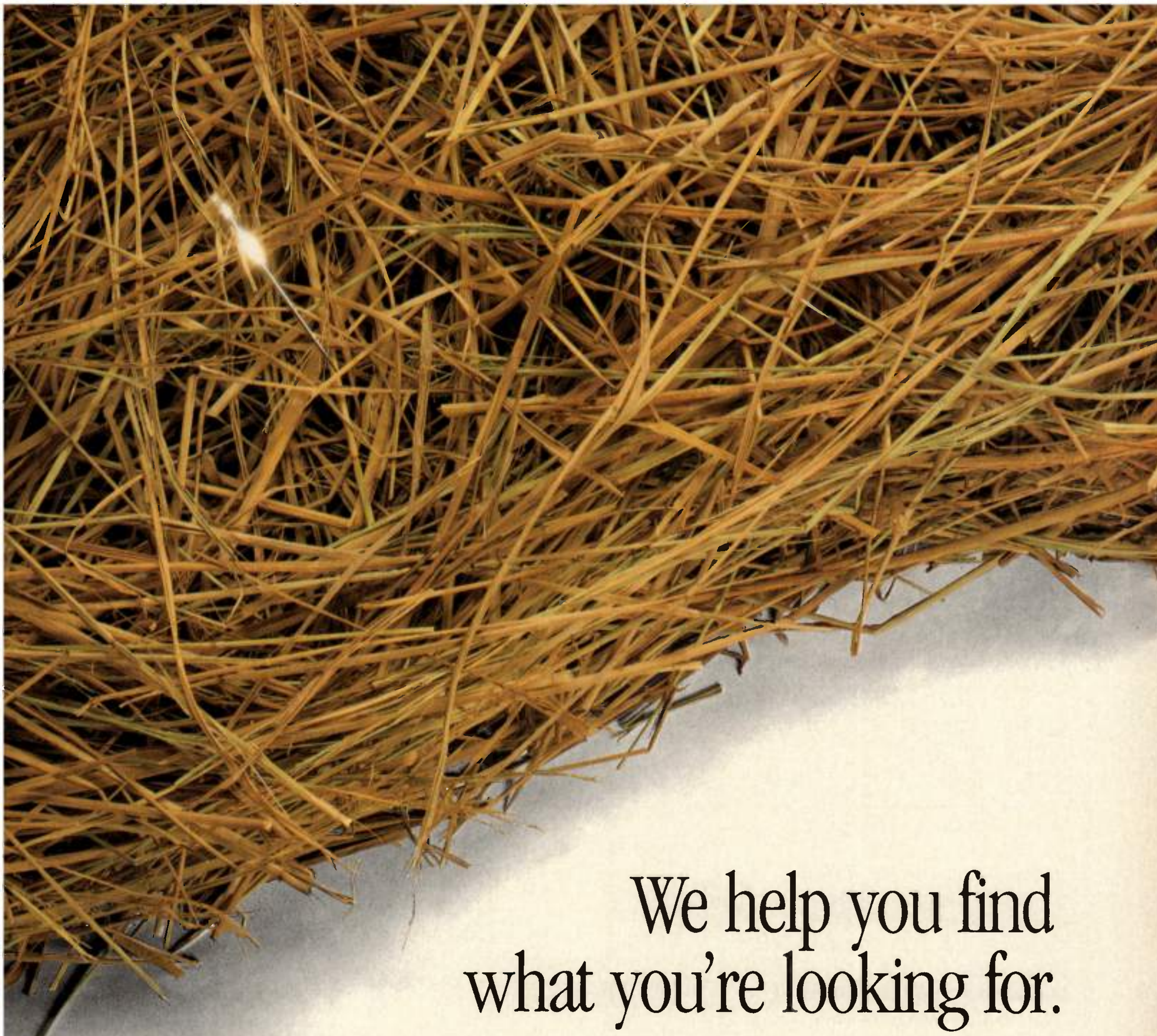

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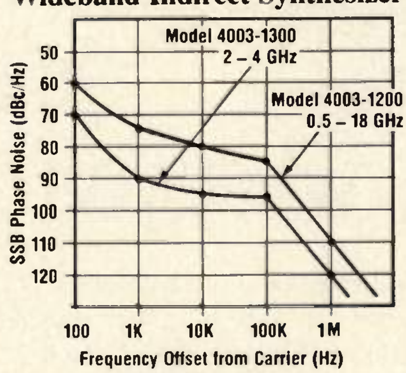

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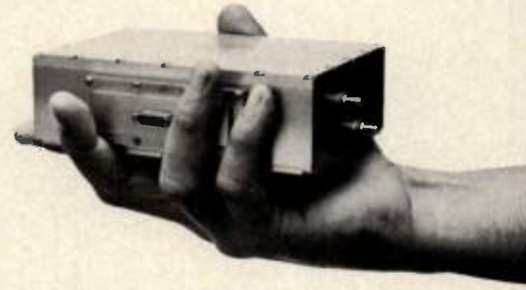

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INFO/CARD 41

Specifying SAW Bandpass Filters

By Lisa Schwartz
Sawtek, Inc.

Complex interdependencies among surface acoustic wave (SAW) performance parameters make it difficult to describe the design limitations and performance options of SAW filters. As with all filters, these interdependent parameters cannot all be optimized simultaneously (Appendix 1). A common pitfall for a systems engineer seeking a SAW filter is the tendency to think in terms of the characteristics of other technologies, such as LC filters. Despite some appealing analogies, SAW specifications for bandpass filters are unique.

The first steps involved in specifying SAW bandpass filters are identifying a 1 dB or 3 dB bandwidth and choosing a center frequency. Figure 1 shows the frequency response parameters. Currently, filters with center frequencies of less than 500 MHz are routinely implemented. Fabrication and photomask difficulties reduce achievable performance when the center frequency of a filter exceeds 500 MHz. However, this does not mean that filters over 1 GHz cannot be produced.

Once these two key parameters are established, the fractional bandwidth (percent BW) can be calculated. This is defined as the 3 dB bandwidth divided by the center frequency and multiplied by 100. The primary design decisions that are necessary to determine if a SAW device will fulfill a given set of requirements are governed by the fractional bandwidth. It is the fractional bandwidth that dictates both the substrate material used and the limitations associated with it. (Table 1 identifies the substrate material most often used with each optimum fractional bandwidth.)

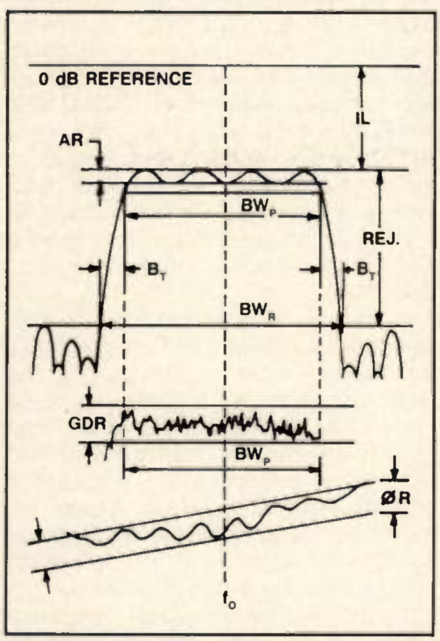

Figure 1. SAW transversal bandpass filter frequency response parameters.

Filters with fractional bandwidths of less than 0.1 percent are too narrow to be implemented as transversal filters. However, such a filter may be ideally suited for SAW resonator filter technology. A resonator filter in this range can be built on quartz for good temperature stability and, in most cases, will exhibit low insertion loss (2 to 6 dB). Near-in rejection of 40 dB is usually attainable with this approach.

A difficult area for SAW designers is that in which filters with fractional bandwidths between 0.1 and 0.25 percent are required. The range is slightly too wide for resonator filters and too narrow for high-performance transversal SAW filters.

For filters with bandwidths up to approximately the 0.5 percent range, a third harmonic design approach will generally be used. This technique usually makes use of the SAW filter response at three times the center frequency ($3f_0$) of the transducers. Consequently, the third harmonic quartz filter will exhibit a fundamental spur at one-third the desired frequency of operation, as well as bulk mode responses at $0.533f_0$ and $0.6f_0$. These spurs are approximately equal in bandwidth to the main response and may only be attenuated 10 dB unless matching elements are designed properly to reject these responses. Insertion loss in the range of 18 dB to 25 dB can be expected for most filters of this type.

The 0.5 to 3.5 percent fractional bandwidth range potentially yields the highest performance quartz SAW filters. Insertion losses typically range from 18 to 30 dB, depending on ripple and VSWR requirements. The design approach is fundamental sampling; therefore, a third harmonic response (at $3f_0$) is seen. In addition to the $3f_0$ spur, bulk mode spurs appear at $1.6f_0$ and $1.8f_0$ and are approximately equal in bandwidth to the desired response. Note that impedance matching is important to the spurious suppression.

Two substrate material options exist in the 3.5 to 6 percent fractional bandwidth range. If quartz temperature stability is desired, near-in rejection of 35 to 45 dB will be exhibited for shape factors as low as 1.3:1. This parameter gradually improves in the 55 to 60 dB range as the shape factor is relaxed. Ultimate rejection is even better. Insertion losses in the 22 to 32 dB range should be expected depending on VSWR, ripple, and time spurious specifications. Frequency spurs may be seen at $1.6 \times f_0$, $1.8 \times f_0$, and $3 \times f_0$ and are influenced by the impedance matching.

If better near-in rejection is needed, or the required shape factor is less than 1.3:1, the filter is better suited for a lithium tantalate substrate material with a temperature coefficient of −23 ppm/°C. Insertion loss then lies between 18 and 27 dB, depending on VSWR, ripple, and time spurious specs. The lithium tantalate filter harmonic response at $3f_0$ is the principal frequency spur. In the 6.0 to 9.0 percent bandwidth range, lithium tantalate substrates are almost always chosen.

Filters with fractional bandwidths in

Material	Approximate Surface Wave Velocity	Temperature Dependency	Optimum Fractional Bandwidth for Each Material
ST Quartz	0.124 in/µsec	$\frac{\Delta f}{f} = -\left(\frac{T-T_0}{5.45}\right)^2$	0.1% - 5%
Lithium Tantalate	0.129 in/µsec	23 ppm/°C	4% - 9%
YZ Lithium Niobate	0.134 in/µsec	94 ppm/°C	7% - 30%
128° Lithium Niobate	0.153 in/µsec	72 ppm/°C	15% - 65%

Table 1. SAW transversal bandpass filter materials parameters.

the 9.0 to 25 percent range are best suited for YZ lithium niobate with a −94 ppm/°C temperature coefficient. If this temperature drift is too high, devices in the narrower portion of this bandwidth range can be put on lithium tantalate. Historically, the highest performance SAW devices fall into this area of fractional bandwidths where low shape factors and high rejection are easy to achieve. The 25 to 68 percent range of bandwidths is better suited for 128° lithium niobate with a −72 ppm/°C temperature coefficient. Sampling problems can appear in this fractional bandwidth range that may limit bandwidth tolerance and amplitude ripple specifications.

Any device that is wider than 40 percent bandwidth requires the use of dispersive techniques such as slanted array correlator (SAC) transducers. Insertion losses in this bandwidth range fall between 20 and 40 dB depending on VSWR, ripple, and time spurious requirements.

The next parameter to be specified is the rejection bandwidth. Typically, this is a 40 or 50 dB bandwidth although the 60 dB points are sometimes used. From this information, the shape factor can be calculated. The shape factor of a bandpass filter is determined by dividing the rejection bandwidth by the 3 dB bandwidth. Filters can be built with shape factors as low as 1.1:1 at certain percent bandwidths; however, this choice influences many other performance parameters.

Insertion loss is another key parameter to specify. Transversal bandpass filters can be divided into two categories: conventional SAW filters and low-loss filters. Conventional SAW bandpass filters can achieve insertion losses as low as 15 dB, although operation in the 20 to 25 dB range is typically preferred, while low-loss filters achieve insertion losses in the 3 to 10 dB range.

Despite the low insertion loss of the low-loss filter, the triple transit can still be suppressed as much as 40 to 50 dB. Suppression of the triple transit is the reason for the higher loss in conventional SAW filters. However, a conventional SAW filter can achieve better near-in rejection and steeper shape factors. Low-loss filters are also more expensive to produce than conventional transversal filters. Matching components are normally required to achieve the desired insertion loss. They can either be integrated inside the package or mounted externally on the board. Innovations in low-loss design techniques are continually improving performance as well as lowering non-recurring and unit costs.

Center frequency, passband width, rejection bandwidth, and insertion loss are very important parameters to the SAW designer. In addition to these parameters, there may be others that are not required to design the SAW filter but may be critical to the performance of the customer's system. These parameters include amplitude ripple, phase ripple, group delay ripple, ultimate rejection, and triple-transit suppression.

As in most devices, cost is often a major factor in SAW filter specification. Sometimes the devices cost several times what they should because specifiers with insufficient knowledge demand

For a given set of design parameters, the following will generally hold true:

- A narrowband filter will be longer than a wideband filter.
- A filter with a low (steep) shape factor will be longer than a device with a higher shape factor.
- A longer device will have worse group delay ripple than a shorter device.
- A lower frequency device will be larger than a higher frequency device.
- The lower the specified rejection level, the longer the device.
- The longer the device, the higher the cost.
- A higher insertion loss will yield better amplitude, phase, and group delay ripple.
- A filter with higher insertion loss will have better triple-transit performance.

Appendix 1. Qualitative overview of parameter interaction.

overly difficult or unnecessary requirements. An obvious example is a center frequency or bandwidth tolerance to be met over a temperature range that is tighter than the natural temperature drift of the material. In that case, the device must be placed in an oven to maintain a fixed temperature. Another frequent error is the requirement for very tight peak-to-peak group delay variation which arises as a response to concerns about the nonlinearity exhibited by other types of filters.

Provided that the frequency spectrum of the signal passing through the device is wide compared to the periodicity of the rapid variations in group delay, these rapid variations will not degrade filter performance. They simply reflect the rapid changes in phase that are occurring. Provided that the phase variations are small enough for the desired application, group delay variations are irrelevant to system performance.

Any parameter that is important to the system is important information to the SAW designer. Table 2 lists many of the parameters and their current limitations. Again, the inter-relationship of all these parameters is complex and not all can be optimized simultaneously. rf

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Parameter Definition	Symbol	Current Practical Limits
Center Frequency	f_O	10 - 1000 MHz
Rejection Bandwidth at Given dB	BW_R	(See SF and % BW)
Passband Width at Given dB	BW_P	(See SF and % BW)
Fractional Bandwidth (or BW_P/f_O)	%BW	0.1% - 65%
Shape Factor (or BW_R/BW_P:1)	SF	1.1:1 - 4.0:1
Rejection	REJ	40 - 70 dB
Insertion Loss	IL	Bidirectional: 15 - 35 dB Unidirectional: 3 - 10 dB
Amplitude Ripple	AR	0.1 - 1 dB
Phase Ripple	øR	1° - 10°
Group Delay Ripple	GDR	± 5 nsec minimum
Transition Bandwidth	B_T	200 kHz minimum

Table 2. Practical limitations of SAW bandpass filter parameters.

rf featured technology

Crystal Delay Equalizers

By William B. Lurie
Consultant

In years gone by, filters were required to have a specified amplitude vs. frequency performance, and phase or delay characteristics were rarely considered. However, with the advent of more sophisticated signal processing, it has been recognized that system performance can be degraded by nonlinear phase or non-uniform envelope delay. In high-speed digital data transmission, for example, the Bit Error Rate can be seriously increased. Fortunately, modern methods permit the design of filters possessing more desirable phase and delay.

Improved delay response vs. frequency can be achieved by a number of techniques. A rather complete treatment of the design of high-grade conventional equalizers using coils and capacitors has been published (1), but the extension of this method to much narrower applications, requiring quartz crystals, has not been addressed. This article presents one of several methods of designing crystal filters with linear phase or constant delay.

The basic technique of equalization is to determine where in the frequency range of interest the filter (or system) has valleys of delay, and then to add one or more equalizer "sections" to fill in those valleys. Typically, a delay-corrector section is an allpass constant-impedance network, having phase nonlinear in such a way as to provide a peaked curve of delay vs. frequency. This is illustrated in Figure 1, where A might be the delay curve for some bandpass filter, B is the delay curve for a typical delay equalizer, and C is that of the combined filter and equalizer.

Very frequently, the delay variation is such that it cannot be adequately flattened over the frequency band of interest using only one equalizer section. A number of similar sections are then used, often staggering the response steepness and peak frequencies. Figure 2 illustrates this approach for the filter in Figure 1 (curve A), using three equalizers. In Figure 2, the delays of the three equalizers are shown as B, C, and D, and the composite delay as E. Note that the total delay has been increased, but the delay variation is reduced considerably over that obtained with only one equalizer.

The synthesis of the equalizer sections is simple and straightforward, and has been well-covered in the article referenced. A typical case is illustrated in Figure 3, which presents the schematic of one of the equalizer sections (the middle frequency of Figure 2, curve B), with element values, for an impedance level of 1000 ohms. As is easily seen, all element values are reasonable and practical.

When very narrow bandpass filters are required, they are frequently built with quartz crystals as resonators. The same principles apply when designing the equalizers, but difficulties arise in attempting to realize the equalizers with practical coils and capacitors. They are obviously narrowband devices, and unless they can be built with quartz crystals possessing high Q, they will be extremely lossy, in addition to developing other problems. No simple variation of the schematic of Figure 3 is available — no canonic form in which crystals can be substituted for a group of coils and capacitors. Fortunately, a somewhat different approach is available, in which crystals can be used to advantage. The theory is outlined below, without going into mathematical detail.

Consider a lossless lattice section, consisting of branches whose reactances are characterized by two sets of

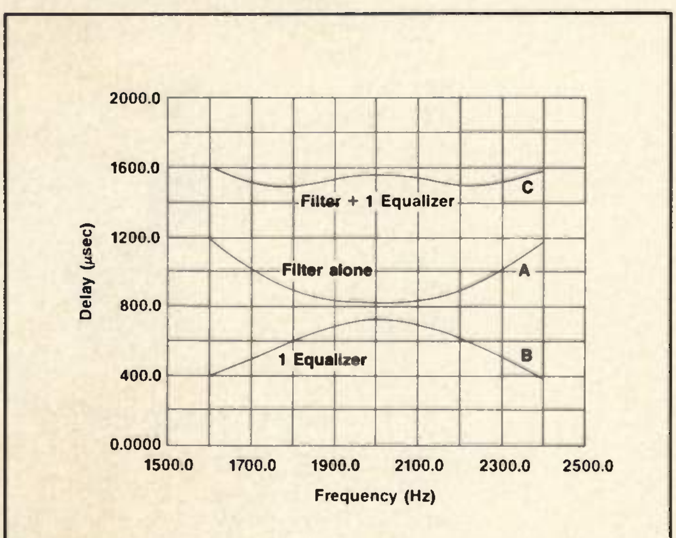

Figure 1. Bandpass filter alone and with one equalizer.

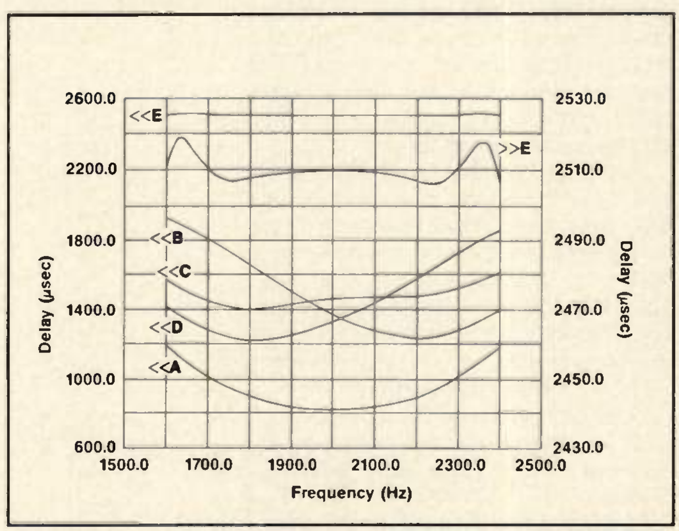

Figure 2. Bandpass filter alone and with three equalizers.

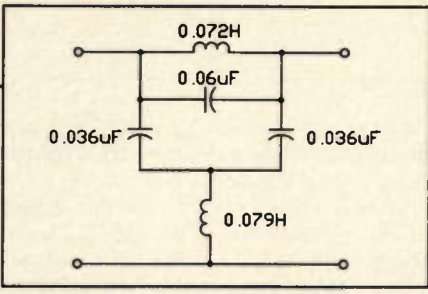

Figure 3. Schematic of the equalizer section of the middle frequency of Figure 2, curve B.

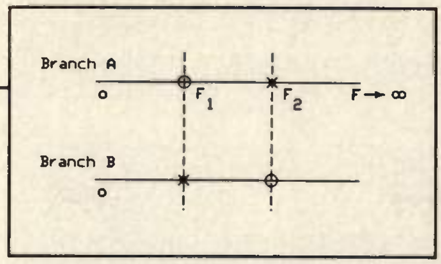

Figure 4. Lossless lattice section, consisting of branches whose reactances are characterized by two sets of poles (X) and zeros (0).

poles and zeros, as shown in Figure 4. Here, an "O" is a frequency at which the branch reactance is zero, and an "X" is a frequency at which the reactance is infinite (a "pole").

Simple network theory states that this describes an allpass network, as the two reactances are reciprocally related. The phase, however, changes 180 degrees as the frequency progresses from F_1 to F_2. The envelope delay for such a section peaks at a frequency about halfway between F_1 and F_2 (for narrow-band cases), and has half that much delay at F_1 and F_2. The actual value of the peak delay is inversely proportional to the separation (F_2-F_1). This gives the capability of building equalizer sections at any peak frequency, with any desired peak value, provided the network can be synthesized in a suitable configuration.

Fortunately, this is quite possible, if use is made of a few tricks of the trade. Branch A is straightforward; it is the usual pole-zero pattern for a quartz crystal, as shown in Figure 5. L_1 and C_1 resonate at F_1, and the anti-resonance is at F_2. Branch B, however, although treated similarly, yields Figure 6, where the shunt capacitance is now negative, since the zero is at a higher frequency than the pole.

In order to transform this into a practical circuit, a way must be found to absorb this awkward component. In practice, it is usually possible to cascade the equalizer with another section or group of elements from which some positive capacitance can be "borrowed."

Suppose, for example, that the complete equalizer section is drawn in cascade with other sections, with element values as indicated. The result is given in Figure 7. Half-lattices are shown, since crystal filters are more frequently built that way.

Figure 7 shows several inductors to ground, each of which resonates at the filter's center frequency. Each resonator's impedance at resonance is large enough to have a negligible effect on the performance of the system. By conventional lattice manipulation methods, 8 pF is borrowed from each of the 50 pF capacitors shown, and is caused to appear instead, inside the equalizer, as 4 pF across the A branch, and 4 pF across the B branch. The result, after combining the +4 pF with the −2 pF capacitance in branch B, is as shown in Figure 8, where a ratio of 250 between the static and motional capacitances of each crystal is assumed. In many cases it is not necessary to add a resonant circuit in order to have capacitance to borrow. For example, in Figure 8 the inductor L_x might be the primary inductance of the "ideal" transformer shown, which might be a necessary part of the crystal lattice section to the left (not drawn). The inductor L_y might be a transformer or autotransformer needed to match the filter to some desired load. In all crystal filter structures of suitable bandwidth, there are capacitors to ground at every node. There is even a monolithic crystal equivalent to the equalizer shown, which has no "ideal" or

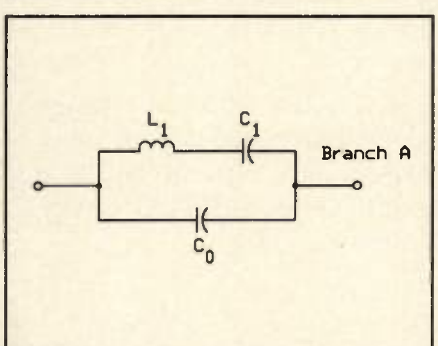

Figure 5. Branch A.

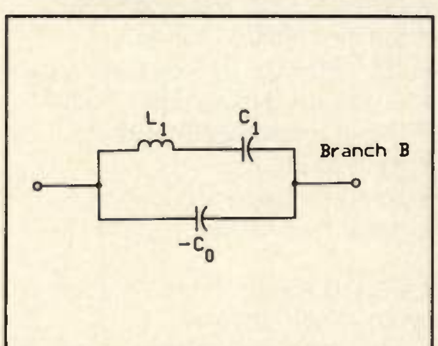

Figure 6. Branch B.

"hybrid" transformer, just as there is a monolithic equivalent to most narrow-band crystal bandpass filters.

As mentioned at the start, this is one of several methods of designing crystal filters with linear phase or constant delay. Some others will be the subjects of future articles.

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William Lurie has worked as a mathematician, physicist and electronics engineer in a variety of fields, including magnetic compasses, X-ray tubes and measuring equipment, airborne Doppler radar, and filters. He holds Life Senior Member status in the IEEE. He works as an independent consultant and can be reached at 5795 Parkwalk Drive, Boynton Beach, FL 33437. The telephone number is (407) 482-8842.

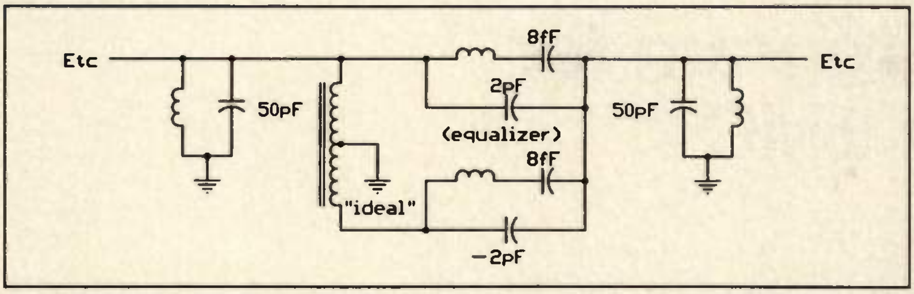

Figure 7. The complete equalizer section drawn in cascade with other sections.

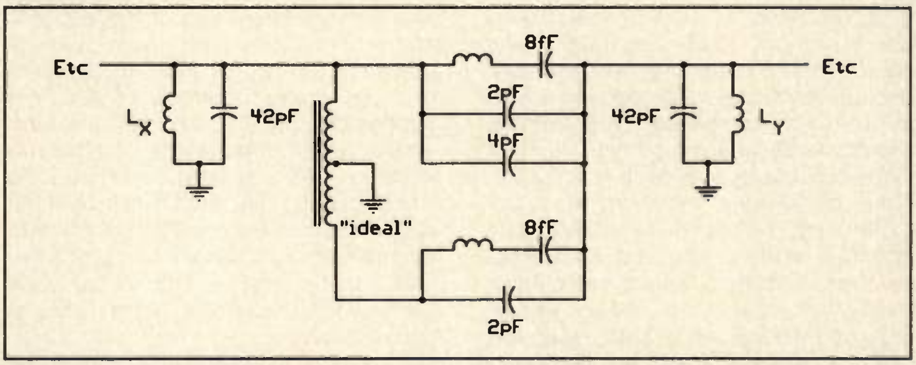

Figure 8. The complete equalizer section after lattice manipulation.

The Navy's Program for Excellence in EMC

A Behind-the-Scenes Look

By James Whalen, SUNY, and
Richard Ford, Naval Research Laboratory

Electromagnetic compatibility (EMC) is an extremely broad-based field dealing with issues as diverse as the effects of lightning striking a shuttle booster, or whether living under power lines is a health hazard. The largest of the subsets of EMC is electromagnetic interference (EMI), when an electronic device can't function properly due to unintended EM energy. A proliferation of new electronic applications, combined with increased circuit sensitivity and complexity, have caused a dramatic increase in EMI.

Because the military is typically in the forefront in applying new technology, there has been a dramatic growth in DOD's need for EMC engineering in a product's life cycle, from design through deployment. Unfortunately, there is a critical shortage of qualified EMC engineers and technicians to assist in assuring EMC. Furthermore, EMC theory is seldom taught as a part of any engineering curriculum and courses on related subjects do not emphasize basic EMC topics. An example is the conventional university course on electromagnetic theory. Far-field radiated components may be treated in detail, but near-field components may be ignored. EM coupling phenomena are seldom discussed. Yet, EMC engineers routinely are concerned with near-field environments and EM coupling modes. So, a BSEE degree alone does not qualify an engineer as an EMC specialist.

Traditionally, the EMC community has been populated by small groups of dedicated specialists in lightning, electromagnetic pulse, electrostatic discharge, and at least a dozen other areas. They are sometimes super-specialists with a sub-specialty within a specialty. They are often few in number and don't effectively interact across the EMC discipline, having their own symposia, and their own terminology. With the increased emphasis on EMC engineering as well as "whole-systems engineering," people supporting these EMC programs have been taxed to their limit.

In response to growing complaints about lack of EMC, both DOD and non-DOD EMC funding over the past decade has dramatically increased. This in turn has stimulated programmatic growth. In the ensuing growth, need for technical quality yielded to quantity. Many of the best engineers were promoted into management earlier than normal in their careers; specialists were made team leaders, sometimes diluting their capabilities in the face of expanding expectations. The end result has been a decrease in the overall quality of EMC technical support.

Other evolutions in the federal system have occurred that further complicate the ability to obtain quality support, with an especially adverse effect on excellence in non-traditional "high tech" areas such as EMC. The Competition in Contracts Act has made it extremely difficult to screen out less-than-capable bidders for engineering service contracts because of the currently acceptable substitutions for engineering or technician training and experience. A common approach is to determine which proposals meet "the minimum requirements" (not who is best), then award the contract to the lowest bidder in that group. Without an established standard for qualifying EMC training and experience, there is no way to stop technically deficient substitutions.

The Electromagnetic Environmental Effects (E3) branch of the Product Integrity Division at the Naval Air Systems Command (NAVAIR) has life-cycle responsibility for aircraft, avionics, electrically initiated air-launched ordnance, and ground support equipment. A portion of that responsibility includes establishment and maintenance of adequate technical resources to ensure that air weapon systems can meet their mission requirements. Early in 1987, in consultation with the Chief of Naval Operations (OP 941F), et al., it was decided that two programs should be evaluated — one for certifying EMC personnel providing direct technical support, the other for accrediting laboratories doing MIL-STD 462 acceptance testing. Certification of personnel to a recognized standard provides a basis for rejecting offers from contractors who do not have experienced, qualified personnel. It also provides a demonstrable benchmark to which government activities may train or replace professional technical staff. Accreditation of laboratories, both public and private, assures that acceptance testing will be accomplished to a consistent standard.

To maximize potential benefit to the technical community for the tax dollars expended on these two programs, both programs have been based outside of DOD. After considerable research, the National Association of Radio and Telecommunication Engineers (NARTE) was selected and agreed to be the agent for certifying EMC technologists. NARTE is a non-profit technical association which has certified over 7000 engineers and technicians in the radio-telecommunications field since 1983. It is important to recognize that NARTE is the administrator of the certification program, while the technical requirements were established by the NAVY/NAVAIR panel. For the long term, an ad hoc committee has been established within the growing ranks of NARTE-certified EMC technologists to provide a framework for EMC technical excellence within NARTE. The National Institute for Standards and Technology (NIST), under its congressionally sponsored National Voluntary Laboratory Accreditation Program (NVLAP), agreed to underwrite MIL-STD 462 laboratory accreditation.

EMC Certification

The certification program began with a group of senior Navy EMC engineers who were deeply concerned about:

1. Improving qualifications of EMC engineers assigned to Navy projects, through education and training.
2. Verifying a minimum level of competency.

3. Certifying a knowledge of the fundamentals of EMC.

4. Providing recognition so that this talent would less likely be lost to the very EMC community which nourished and sustained it.

The initial group was soon expanded to include others from government, industry, and universities. Under the tutelage of Russ Carstensen, head of NAVAIR 5161, open-to-all monthly meetings were held in Washington, D.C. starting in the fall of 1987. Over a period of about five meetings the panel established the baseline of the certification program. Certification of personnel would be a four-step process based on education, work experience, peer endorsement and examination. It would bear resemblance to both the states' Professional Engineer's licensing programs and the IEEE membership process (1).

The panel quickly recognized the need to develop a bank of examination questions. This bank would be a key element in the procedure for certifying the competency of EMC technologists. It would also be an invaluable contribution to EMC technology by formally documenting the basics of EMC. Towards this end, the panel saw the need to solicit exam questions from the members of the EMC community itself. Hence, all grandfather candidates have been required to submit a list of ten questions and answers. Over a thousand questions have been submitted and the number grows daily.

The panel also recognized the need to have a quality exam, with good questions, clearly stated and unambiguously arranged. For this, Dr. James Whalen was invited to join the panel. Not only does he have solid EMC credentials, but he has made a personal hobby of studying the essential criteria for quality multiple-choice exams in the technical area. He was appointed to head the sub-group responsible for the exam. The primary role of this group is to evaluate, groom and clarify the submitted exam questions. Particularly troublesome areas are ambiguous meaning and unconventional terminology. Often, two group members feel they clearly understand the question, but when they get together they realize that each understands it differently!

Some have expressed concern about EMC certification becoming a process by which an elite group is certified. The exam group's goal is to produce an exam that at least two thirds of the candidates would pass, a major challenge given the broad coverage of EMC. The exam contains both published and unpublished questions, and is open book in the freest sense (we call it the "one wheel-barrow exam"). Although unlikely to be confirmed officially, the first few exams will likely err on the side of caution; they might be easy. This is compatible with the goal of slowly but surely raising the technical quality of the community and validating it with tougher future exams.

NVLAP Accreditation

The Navy's program for EMC quality also requires laboratories doing MIL-STD 462 final acceptance testing to be accredited by NVLAP (2), in order to bid on work for contracts issued by NAVAIR after November 1989. In addition, the quality assurance provisions of most DOD (as well as some non-DOD) work, argue strongly the value of NVLAP accreditation.

One of the major behind-the-scenes issues is whether this NVLAP will have the same rough going as had the FCC NVLAP. This is very unlikely. The hard requirement in NAVAIR 2410.1D for NVLAP means that people on both sides of the table must be responsive to each other. NVLAP's policy is to rely wholly on non-staff (non-NIST employees) as technical assessors to provide the technical backbone of the program. As long as these assessors come from the ranks of DOD's MIL-STD 462 corporate knowledge (which is very much the present case), the program is unlikely to be anything but healthy.

In June, 32 candidate assessors for the 462 NVLAP completed training. They will be assessing a model lab in the fall and assessing applicant labs shortly thereafter. As in the certification program, there's strong emphasis on not creating any artificial shortage of available labs. The guidance is to resolve any deficiencies by working cooperatively with the applicants, recognizing that each lab has declared its desire and commitment to adhere to NVLAP's standards. Concern comes from many in the EMC community who feel there are some severe shortcomings to MIL-STD 461/462. There are high hopes that the artifact testing, round robins, and sponsor/applicant lab interactions will focus productive energy on resolving these issues. We believe over the long run such improvements will come, but must point out that the DOD agencies responsible for these specs are in two different services and have no formal relation to either NVLAP or the assessors (and perhaps even to each other).

A Program for EMC Excellence

The newly released NAVAIR INSTRUCTION 2410.1D dated May 17, 1989 (3) mandates that personnel in responsible charge of EMC work must be NARTE EMC certified and that MIL-STD 462 testing for record may only be done by NVLAP-accredited laboratories. Both programs are autonomous from the Navy and are user-funded, requiring no "after start-up" tax dollars.

Certification and accreditation are two separate parts of the Program for EMC Excellence. They have no common elements except they both address EMC technology. Accredited labs do not have to have certified technologists! (Prudent labs will likely see the value in employing candidates who have taken the time to validate their credentials.) Each program has a life of its own, one in a non-profit private organization, the other in a non-DOD government agency. Maximum effort has been made to base their future in their own hands. And maximum effort has been made to not color them Navy blue (or Army green, etc.), although the NARTE certificates are red, white and...blue. rf

References

1. The exact requirements are given in the "EMC Credential Certification Handbook" available from NARTE, P.O. Box 15029, Salem, OR 97309, Tel: (503) 581-3396.
2. Details and application forms for the 462 NVLAP are available from NIST, NVLAP Bldg 411, RM A124, Gaithersburg, MD 20899, Tel (301) 975-4016.
3. Order NAVAIR 2410.1D from Naval Publications and Forms Center, 6801 Tabor Ave., Philadelphia, PA 19120-5099.

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INFO/CARD 51

Chebyshev Filters With Arbitrary Source and Load Resistances

By Jack Porter
Cubic Corporation

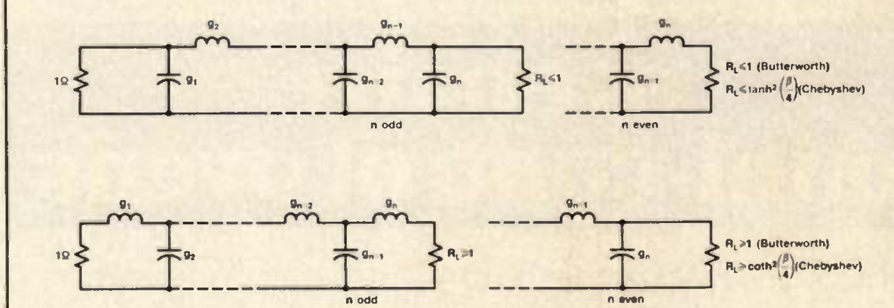
The prototype element values required for passive filter design are usually obtained from tables (1), but these tables are far from complete. Tabulated values are available for only a limited number of load-to-source resistance ratios and, for Chebyshev filters, only a few values of passband ripple. Fortunately, values can be calculated quite easily for a Chebyshev or Butterworth filter with any resistance ratio and ripple. This article describes a computer program which performs all of the calculations required to design Chebyshev or Butterworth lowpass or highpass filters.

Simple recursion formulas for the prototype element values of singly terminated filters, and for those of doubly terminated filters with minimum insertion loss, were first published many years ago and are well-known (2,3). Although explicit formulas for doubly terminated filters with any ratio of load-to-source resistance have been derived by Takahasi (4,5) and by Rhodes (6), they are in forms which do not permit easy calculation of the element values. However, the recursion formulas shown in Figure 1 can be derived from Rhodes' formulas and, with somewhat more difficulty, from Takahasi's (7).

Table 1 contains definitions of common Chebyshev filter parameters. Figure 1 shows the various doubly terminated prototype filter configurations along with the applicable formulas. The value of A_t calculated by these formulas is the insertion loss expressed as a power ratio for an ideal Butterworth or odd-order Chebyshev filter. For an even-order Chebyshev filter, it is the insertion loss in excess of the required minimum.

For odd-order filters with $R_L \leq 1$, there are shunt capacitors at both ends, while in the dual filter with $R_L \geq 1$ there are series inductors at both ends. Since the filter is reciprocal (the source and load can be interchanged), there are these two possible configurations for any odd-order filter.

Even-order filters have a series inductor at the low resistance end and a shunt



Chebyshev:

$$A_t = \frac{4R_L}{(R_L+1)^2} \quad \text{(n odd)} \qquad A_t = \frac{4R_L A_p}{(R_L+1)^2} \quad \text{(n even)}$$

$$\gamma = \sinh\left(\frac{\beta}{2n}\right) \qquad \delta = \sinh\left[\left(\frac{1}{n}\right)\sinh^{-1}\left(\frac{\sqrt{1-A_t}}{\epsilon}\right)\right]$$

$$b_R = \gamma^2 + \delta^2 - 2\gamma\delta\cos\left(\frac{R\pi}{n}\right) + \sin^2\left(\frac{R\pi}{n}\right) \qquad R = 1,2,\ldots,n$$

Butterworth:

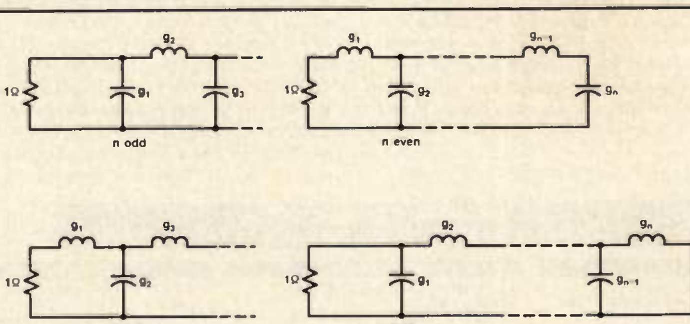
$$A_t = \frac{4R_L}{(R_L+1)^2}$$

$$\gamma = 1 \qquad \delta = (1 - A_t)^{1/(2n)}$$

$$b_R = \gamma^2 + \delta^2 - 2\gamma\delta\cos\left(\frac{R\pi}{n}\right) \qquad R = 1,2,\ldots,n$$

$$a_R = \sin\left[\frac{(2R-1)\pi}{2n}\right] \quad R = 1,2,\ldots,n \qquad g_1 = \frac{2a_1}{\gamma - \delta} \qquad g_R = \frac{4a_R a_{R-1}}{b_{R-1}g_{R-1}} \quad R = 2,3,\ldots,n$$

Figure 1. Doubly terminated filters.



Chebyshev:

$$\gamma = \sinh\left(\frac{\beta}{2n}\right)$$

$$b_R = \left[\gamma^2 + \sin^2\left(\frac{R\pi}{2n}\right)\right]\cos^2\left(\frac{R\pi}{2n}\right) \qquad R = 1,2,\ldots,n$$

Butterworth:

$$\gamma = 1$$

$$b_R = \cos^2\left(\frac{R\pi}{2n}\right) \qquad R = 1,2,\ldots,n$$

$$a_R = \sin\left[\frac{(2R-1)\pi}{2n}\right] \qquad R = 1,2,\ldots,n \qquad g_1 = \frac{a_1}{\gamma}$$

$$g_R = \frac{a_R a_{R-1}}{b_{R-1}g_{R-1}} \qquad R = 2,3,\ldots,n$$

Figure 2. Singly terminated filters.

capacitor at the high resistance end in both cases. For minimum insertion loss filters with the load resistance equal to the maximum or minimum possible value, the component values are the same in both cases. For other ratios of load-to-source resistance, however, there are two completely different sets of element values.

The singly terminated prototypes and formulas are shown in Figure 2. The final prototype element is always either a capacitor with an open circuit filter termination or an inductor with a short circuit termination.

The noise bandwidth of any type of all-pole filter can be calculated easily if its singly terminated prototype element values are known (8). It has been shown (9) that the relative noise bandwidth of these filters is $(\pi/2)/C$, where C is the value of the capacitor adjacent to the open circuit in the singly terminated prototype (i.e., $C=g_n$ in this case). For Chebyshev filters, this is the noise bandwidth relative to either the ripple bandwidth or the 3 dB bandwidth, depending on which prototype values were calculated.

The Chebyshev prototype element values calculated using the formulas in Figures 1 and 2 are for filters with a 1 radian/s ripple bandwidth. For a 1 radian/s 3 dB bandwidth filter, each value must be multiplied by the parameter ω_h, which is calculated using the formulas in Table 1. (It should be noted that the 3 dB bandwidth is defined by Zverev (1) as the point which is 3 dB down from the ripple peaks, not from the response at zero frequency. In other words, the formula given in Table 1 for ω_h with n even is also used for n odd. These values will be calculated by the program in Table 2 if line 1000 is deleted.) The Butterworth prototype values are for a 1 radian/s 3 dB bandwidth.

After the prototype element values have been found, the actual lowpass or highpass filter component values are calculated by frequency and impedance scaling. For a filter cutoff frequency F_c (in Hertz) and $\omega_c = 2\pi F_c$, the values are calculated as follows: For lowpass filters, each capacitor in the prototype becomes a capacitor of value $(g_R)/(\omega_c R)$. Each inductor in the prototype becomes an inductor of value $(Rg_R)/(\omega_c)$. For highpass filters, each capacitor in the

n: number of poles or filter elements
A_m: passband ripple in dB

$$A_p = 10^{(A_m/10)}$$

$$\varepsilon = \sqrt{A_p - 1}$$

$$\beta = 2\sinh^{-1}\left(\frac{1}{\varepsilon}\right)$$

$$\gamma = \sinh\left(\frac{\beta}{2n}\right)$$

3 dB (half-power) bandwidth:

$$\omega_h = \cosh\left[\left(\frac{1}{n}\right)\cosh^{-1}\left(\frac{\sqrt{A_h - 1}}{\varepsilon}\right)\right]$$

where $A_h = 2$ for n odd
$A_h = 2A_p$ for n even

Table 1. Chebyshev filter parameter definitions.

```
100 'Program CBLCF
110 'Definitions of variables:
120 'Am: Passband ripple in dB (Chebyshev filters);
130 '    for Butterworth filters Am=0.
140 'C's & L's: Component values in farads & henries.
150 'Fc: Filter cut-off frequency in Hz.
160 'N: Number of poles.
170 'RS & RL: Source & load resistances in ohms.
180 DEFINT I-K, N
190 DIM A(12), C(12,2), C$(2), D$(2), E$(2), F$(4), G(12), I(16,2)
200 PI=3.141593: DP=10/LOG(10)
210 F1$="RS=1   RL zero or infinite"
220 F2$="RS=1   RL=##.####"
230 F3$=" or #.#####"
240 F4$="Insertion loss=##.## dB"
250 F5$="Prototype element values:"
260 F6$="##.####     "
270 F7$="!##=##.####^^^^"
280 C$(1)="L": C$(2)="C"
290 D$(0)="Prototype ": D$(1)="Low-pass ": D$(2)="High-pass "
300 E$(0)="Butterworth ": E$(1)="Chebyshev ": E$(2)=E$(1)
310 F$(1)="RS=\     \            RL infinite"
320 F$(2)="RS=\     \            RL=0"
330 F$(3)="RS infinite           RL=\     \"
340 F$(4)="RS=0                  RL=\     \"
350 FOR K2=1 TO 2: FOR K1=1 TO 16: READ I(K1,K2): NEXT K1, K2
360 DATA 2, 2, 2, 1, 1, 1, 1, 2, 1, 1, 1, 2, 2, 2, 2, 1
370 DATA 1, 2, 2, 1, 2, 1, 2, 1, 2, 1, 1, 2, 1, 2, 1, 2
380 DEF FNSN(X)=(EXP(X)-EXP(-X))/2                     'sinh
390 DEF FNCS(X)=(EXP(X)+EXP(-X))/2                     'cosh
400 DEF FNCT(X)=1+2/(EXP(2*X)-1)                       'coth
410 DEF FNAS(X)=LOG(X+SQR(X*X+1))                  'arc sinh
420 DEF FNAC(X)=LOG(X+SQR(X*X-1))                  'arc cosh
430 CLS: PRINT "CHEBYSHEV & BUTTERWORTH L-C FILTERS"
440 PRINT "Filter types:  1. Low-pass  2. High-pass"
450 PRINT: INPUT "Type, N, Am(dB)"; I1, NP, AM
460 IF I1<=0 THEN 2110
470 INPUT "1. Doubly terminated  2. Singly terminated"; I2
480 ON I2+1 GOTO 430, 490, 550
490 INPUT "RS, RL"; R1, R2
500 IF R1=0 THEN 450
510 R0=R2/R1                                   'RL (prototype)
520 T1=R0+1: AT=4*R0/(T1*T1)                               'At
530 AL=-DP*LOG(AT)                             'Insertion loss
540 IF R0<1 THEN I6=1: R0=1/R0 ELSE I6=2
550 I5=NP-2*INT(NP/2)
560 FOR K1=1 TO NP
570 A(K1)=SIN((2*K1-1)*PI/(2*NP))                          'a(k)
580 NEXT K1
590 IF AM>0 THEN 630
600 I3=0: G1=1
610 IF I2=1 THEN D1=(1-AT)^(1/(2*NP))                 'delta (B)
620 ON I2 GOTO 780, 880
630 INPUT "Specify:  1. Ripple BW, 2. 3 dB BW"; I3
640 AP=EXP(AM/DP)                                          'Ap
650 E1=SQR(AP-1)                                      'epsilon
660 B2=FNAS(1/E1)                                      'beta/2
670 G1=FNSN(B2/NP)                                      'gamma
680 IF I2=2 THEN 880
690 IF I5>0 THEN 760
700 AT=AT*AP
710 IF AT<=1 THEN 760
720 PRINT "Filter not realizable with RL="; R2
730 AT=1: AL=AM: T1=FNCT(B2/2): R0=T1*T1
740 IF I6=1 THEN R2=R1/R0: PRINT "Maximum"; ELSE R2=R1*R0: PRINT "Minimum";
750 PRINT " value of RL="; R2
760 T1=FNAS(SQR(1-AT)/E1)
770 D1=FNSN(T1/NP)                                  'delta (C)
780 G(1)=2*A(1)/(G1-D1)                                  'g(1)
790 FOR K1=2 TO NP
800 T1=(K1-1)*PI/NP
810 B1=G1*G1+D1*D1-2*G1*D1*COS(T1)                'b(k-1) (B)
820 IF I3=0 THEN 850
830 T2=SIN(T1)
840 B1=B1+T2*T2                                   'b(k-1) (C)
850 G(K1)=4*A(K1-1)*A(K1)/(B1*G(K1-1))                   'g(k)
860 NEXT K1
870 GOTO 980
880 G(1)=A(1)/G1                                         'g(1)
890 FOR K1=2 TO NP
900 T1=(K1-1)*PI/(2*NP)
910 T2=COS(T1)
920 B1=T2*T2                                      'b(k-1) (B)
930 IF AM=0 THEN 960
940 T2=SIN(T1)
950 B1=B1*(G1*G1+T2*T2)                           'b(k-1) (C)
960 G(K1)=A(K1)*A(K1-1)/(B1*G(K1-1))                     'g(k)
970 NEXT K1
980 IF I3<2 THEN 1040
990 AH=2                                                   'Ah
1000 IF I5=0 THEN AH=AH*AP
1010 T1=FNAC(SQR(AH-1)/E1)
1020 WH=FNCS(T1/NP)                                        'Wh
1030 FOR K1=1 TO NP: G(K1)=WH*G(K1): NEXT K1
1040 IF I2=2 THEN PRINT F1$: GOTO 1080
1050 PRINT USING F2$; R0;
1060 IF R0>1 THEN PRINT USING F3$; 1/R0 ELSE PRINT
1070 PRINT USING F4$; AL
1080 PRINT F5$
1090 FOR K1=1 TO NP: PRINT USING F6$; G(K1),
1100 IF K1=5*INT(K1/5) THEN PRINT
1110 NEXT K1: PRINT
1120 I4=0: GOSUB 1780
1130 IF I2=2 THEN 1460
1140 PRINT: INPUT "Fc"; F0
1150 ON SGN(F0)+2 GOTO 2110, 430, 1160
1160 W0=2*PI*F0
1170 I4=2*I1+I6-2
1180 FOR K1=1 TO NP STEP 2
1190 K2=K1+1: K3=NP-K1+1
1200 ON I4 GOTO 1210, 1270, 1330, 1390
1210 C(K1,1)=G(K1)/(R1*W0)                          'LP, RL<RS
1220 C(K3,2)=G(K1)*R2/W0
1230 IF K2>NP THEN 1440
1240 C(K2,1)=G(K2)*R1/W0
1250 C(K3-1,2)=G(K2)/(R2*W0)
1260 GOTO 1440
1270 C(K1,1)=G(K1)*R1/W0                           'LP, RL>=RS
1280 C(K3,2)=G(K1)/(R2*W0)
1290 IF K2>NP THEN 1440
1300 C(K2,1)=G(K2)/(R1*W0)
1310 C(K3-1,2)=G(K2)*R2/W0
1320 GOTO 1440
1330 C(K1,1)=R1/(G(K1)*W0)                          'HP, RL<RS
1340 C(K3,2)=1/(G(K1)*R2*W0)
1350 IF K2>NP THEN 1440
1360 C(K2,1)=1/(G(K2)*R1*W0)
1370 C(K3-1,2)=R2/(G(K2)*W0)
1380 GOTO 1440
1390 C(K1,1)=1/(G(K1)*R1*W0)                       'HP, RL>=RS
1400 C(K3,2)=R2/(G(K1)*W0)
1410 IF K2>NP THEN 1440
1420 C(K2,1)=R1/(G(K2)*W0)
1430 C(K3-1,2)=1/(G(K2)*R2*W0)
1440 NEXT K1
1450 GOTO 1590
1460 INPUT "1. Source resistor   2. Load resistor"; I4
1470 ON SGN(I4)+2 GOTO 2110, 430, 1480
1480 IF I4=1 THEN PRINT "Fc(Hz), RS";: ELSE PRINT "Fc(Hz), RL";
1490 INPUT F0, R1
1500 W0=2*PI*F0: K2=I1
1510 FOR K1=1 TO NP
1520 K3=NP-K1+1: X1=G(K3)              'Start from unterminated end
1530 IF I1=2 THEN X1=1/X1                                  'HP
1540 IF I4=2 THEN K3=K1
1550 C(K3,K2)=X1/(R1*W0)                 'LP, open or HP, short
1560 K2=3-K2                             'K2: 1. open, 2. short
1570 C(K3,K2)=X1*R1/W0                   'LP, short or HP, open
1580 NEXT K1
1590 FOR K2=1 TO 2
1600 IF I2=2 THEN 1640
1610 J2=8*(I1-1)+4*(I6-1)+2*I5+K2
1620 GOSUB 1690
1630 GOTO 1670
1640 J1=2*(I4-1)+K2
1650 J2=8*(I1-1)+4*I5+J1
1660 GOSUB 1710
1670 NEXT K2
1680 GOTO 430
1690 PRINT: PRINT "RS="; R1; TAB(21) "RL="; R2
1700 GOTO 1720
1710 PRINT: PRINT USING F$(J1); STR$(R1)
1720 J3=1: J4=I(J2,I2): FOR K1=1 TO NP
1730 PRINT TAB(J3) USING F7$; C$(J4); K1; C(K1,K2);
1740 J3=J3+20: IF J3>41 THEN J3=1: PRINT
1750 J4=3-J4
1760 NEXT K1
1770 IF J3>1 THEN PRINT
1780 PRINT: INPUT "Print results"; I$
1790 IF LEFT$(I$,1)="N" THEN 2100
1800 IF I7=1 THEN 1830
1810 LPRINT CHR$(27)+CHR$(108)+CHR$(10);                'Left margin
1820 LPRINT CHR$(27)+CHR$(78)+CHR$(8): I7=1    'Skip over perforation
1830 IF I4=0 THEN J3=I4 ELSE J3=I1
1840 LPRINT CHR$(13); CHR$(14); E$(I3); D$(J3); "Filter"; CHR$(13)
1850 LPRINT "N="; NP;
1860 IF I3>0 THEN LPRINT TAB(21) "Am="; AM; "dB";
1870 IF I3=1 THEN LPRINT TAB(41) "Ripple BW";
1880 IF I3=2 THEN LPRINT TAB(41) "3 dB BW";
1890 LPRINT
1900 IF I4=0 THEN 2010
1910 IF I2=1 THEN LPRINT "RS="; R1; TAB(21) "RL="; R2;
1920 IF I2=2 THEN LPRINT USING F$(J1); STR$(R1);
1930 J3=1: J4=I(J2,I2): LPRINT TAB(41) "Fc="; F0; "Hz"
1940 FOR K1=1 TO NP
1950 LPRINT TAB(J3) USING F7$; C$(J4); K1; C(K1,K2);
1960 J3=J3+20: IF J3>41 THEN J3=1: LPRINT
1970 J4=3-J4
1980 NEXT K1
1990 IF J3>1 THEN LPRINT
2000 GOTO 2090
2010 IF I2=2 THEN LPRINT F1$: GOTO 2050
2020 LPRINT USING F2$; R0;
2030 IF R0>1 THEN LPRINT USING F3$; 1/R0 ELSE LPRINT
2040 LPRINT USING F4$; AL
2050 LPRINT F5$
2060 FOR K1=1 TO NP: LPRINT USING F6$; G(K1),
2070 IF K1=5*INT(K1/5) THEN LPRINT
2080 NEXT K1: LPRINT
2090 LPRINT CHR$(13)
2100 RETURN
2110 IF I7=1 THEN LPRINT CHR$(12)
2120 END
```

Table 2. The filter program listing.

```
Chebyshev Prototype Filter
N= 5                    Am= .1 dB           3 dB BW
RS=1    RL= 5.0000 or 0.20000
Insertion loss= 2.55 dB
Prototype element values:
 7.8890      0.3659      9.1272      0.2950      3.5457

Chebyshev Low-pass Filter
N= 5                    Am= .1 dB           3 dB BW
RS= 50                  RL= 250             Fc= 2000000 Hz
C 1= 1.1286E-09         L 2= 5.8695E-06     C 3= 2.9053E-09
L 4= 7.2798E-06         C 5= 2.5111E-09
```

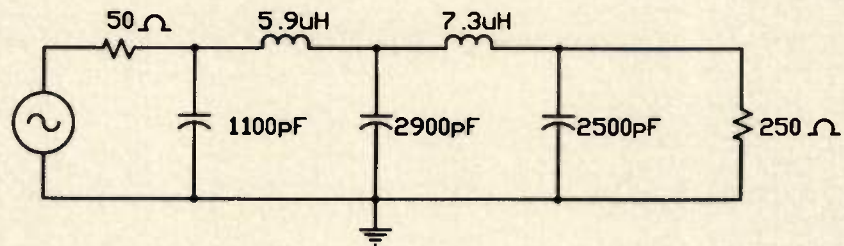
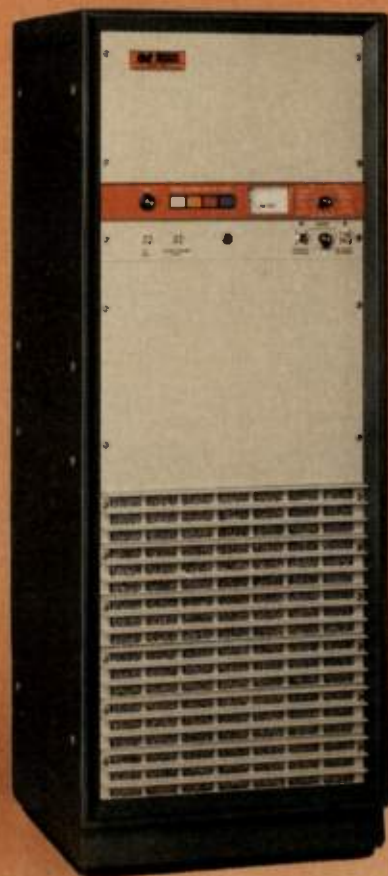


Figure 3. Chebyshev lowpass filter #1.

prototype becomes an inductor of value $R/(\omega_c g_R)$. Each inductor in the prototype becomes a capacitor of value $1/(\omega_c R g_R)$. For these calculations, the "R" can be either the source or the load resistance, whichever corresponds to the 1 ohm resistor in the prototype.

Bandpass filter designs are always based on doubly terminated, minimum insertion loss prototypes. Transformers or other impedance-matching networks are used for any required impedance transformations at the input or the output (3).

Table 2 is a program written in Microsoft Basic which performs all the calculations required to design Chebyshev or Butterworth lowpass or highpass filters. Figures 3 through 5 show examples of filters designed using



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Sound too good to be true? The first two probably are, but $30 a watt is about all you'll pay for our new model 1000L power amplifier—$30 a watt in the bandwidth you need for rf susceptibility testing.

For little more than $30 a watt, the 1000L delivers a full 1000 watts over a bandwidth of 10 kHz to 220 MHz. In pulse mode, the output rating is more than doubled—2500 watts to 150 MHz—or a little under $13 a watt.

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INFO/CARD 55

42809

```
Butterworth Prototype Filter
N= 6
RS=1    RL= 3.3333 or 0.30000
Insertion loss= 1.49 dB
Prototype element values:
 5.2804       0.5517       5.4325       0.3788       2.6559
 0.0816

Butterworth High-pass Filter
N= 6
RS= 1000                RL= 300                 Fc= 1E+07 Hz
L 1= 3.0140E-06         C 2= 2.8849E-11         L 3= 2.9297E-06
C 4= 4.2015E-11         L 5= 5.9924E-06         C 6= 1.9493E-10
```

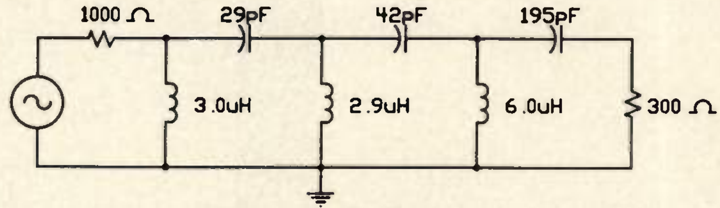

Figure 4. Butterworth highpass filter.

```
Chebyshev Prototype Filter
N= 4                 Am= .5 dB             3 dB BW
RS=1    RL zero or infinite
Prototype element values:
 0.9240      1.5395      1.9116      1.4534

Chebyshev Low-pass Filter
N= 4                 Am= .5 dB             3 dB BW
RS=0                 RL= 1000              Fc= 1000000 Hz
L 1= 2.3132E-04      C 2= 3.0424E-10       L 3= 2.4503E-04
C 4= 1.4705E-10
```

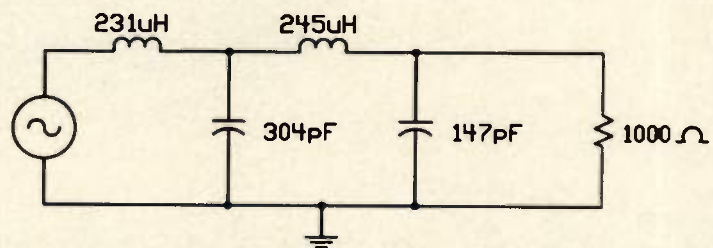

Figure 5. Chebyshev lowpass filter #2.

this program. An HP 15C calculator program which calculates the doubly terminated prototype element values is also available. This program is available on disk from the RF Design Software Service. See page 14 for details. rf

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4. H. Takahasi, "On the Ladder Type Filter Network with Tchebycheff Response," *J. Inst. Elec. Commun. Engrs. Japan*, Vol. 34, pp. 65-74, February, 1951.
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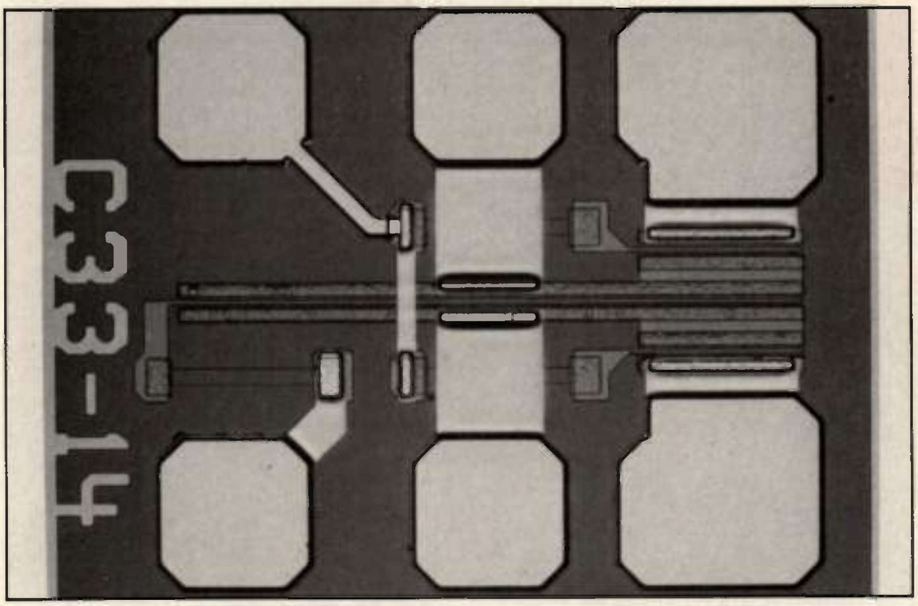
9. W. Ku, "Noise Margin and Gain-Bandwidth Limitations of Legendre Filters," *IEEE Trans. on Circuit Theory*, Vol. CT-17, pp. 670-672, November, 1970.

About the Author

Jack Porter is a senior technical specialist in the Defense Systems Division of Cubic Corporation, 9333 Balboa Avenue, P.O. Box 85587, San Diego, CA 92138-5587 The telephone number is (619) 277-6780.

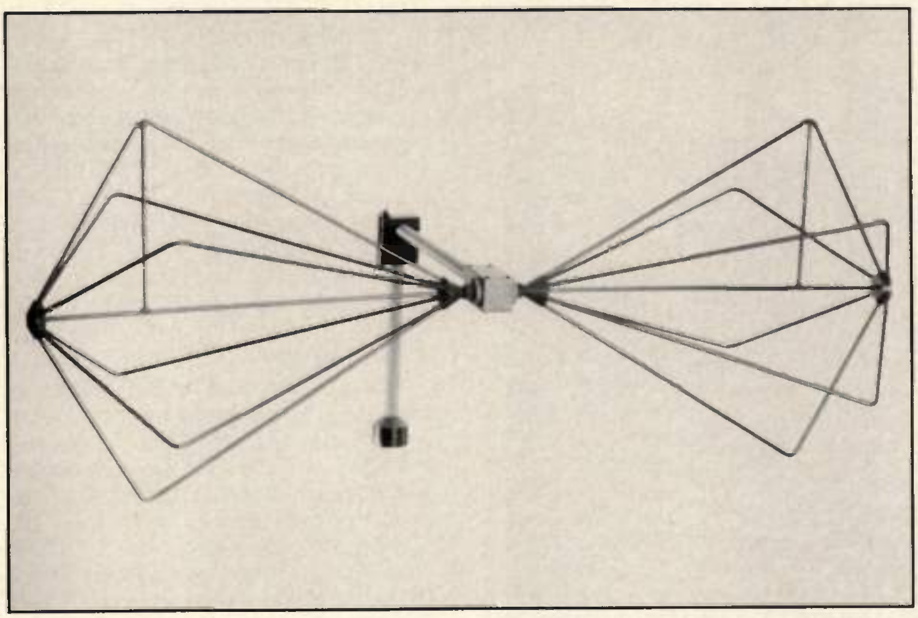
Harris Introduces Cascadable GaAs MMIC Attenuator



The HMR-11000 is a voltage-variable GaAs MMIC attenuator in chip form. Directly cascadable without interstage matching, the device is suitable for a variety of instrument and telecommunications applications from DC to 18 GHz. Typical attenuation ranges from 35 dB at 2 GHz to 15 dB at 18 GHz. From DC to 8 GHz, VSWR is 1.2:1 and input power for 1 dB change in attenuation is 16 dBm. Third-order intercept point is typically 40 dBm. Attenuation switching time for 10 to 90 percent is 3 ns.

The device uses three GaAs FETs in a pi configuration, with insertion loss of 2.0 dB at 12 GHz. The separate input and output ground pads resulting from the pi arrangement improve isolation and gain/phase dispersion. Pricing is $12.30 in 1000s. **Harris Microwave Semiconductor, Milpitas, CA. Please circle INFO/CARD #230.**

EMCO Announces Biconical Antenna



Model 3110 is a biconical antenna that combines balun, feedline and element design to achieve linear response curves. In the balun, a ferrite splits the feedline current for a better balanced-to-unbalanced current transformation. In the feedline, ferrites are used to reduce interference from common mode current.

A new element design eliminates a 281 MHz "bump" found in many biconical antennas. The design also reduces sensitivity to vertical plane orientation, minimizing possible errors. Standard feed connectors are type N. Applications include site attenuation and continuous swept measurements for automated testing facilities. The Model 3110 is individually calibrated and shipped with certified test results. **The Electro-Mechanics Company (EMCO), Austin, TX. Please circle INFO/CARD #229.**

New Logic-Compatible Oscillators

The CXO-63, MXO-53 and VXO-55 have been introduced by CTS Corp. The CXO-63 series offers an H-CMOS compatible, tristatable output over 1 to 70 MHz, and offers a choice of ± 4 mA low drive capability to allow for improved EMI/RFI performance, or ± 16 mA high drive. This series is applicable for use with the Intel 80386 microprocessor.

The MXO-53 series is a true TTL compatible, tristatable oscillator for 200 Hz to 70 MHz. Packaged in a 14-pin hermetic DIP, optional stabilities include ± 25 ppm (0-70°C) and ± 50 ppm (−55 to +125°C).

Also available is the VXO-55 series of TTL compatible, voltage-controlled oscillators, ideal for phase-locked loop applications in the 765 kHz to 35 MHz range. The pullability ranges offered are ± 120 ppm or ± 200 ppm minimum. **CTS Corp./Knights Division, Sandwich, IL. INFO/CARD #228.**

50 Ohm Feedthroughs

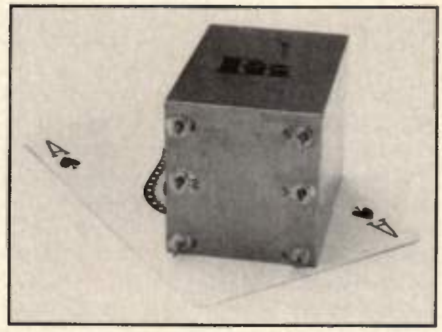
Kyle Technology Corp. introduces 50 ohm impedance, hermetically sealed, aluminum RF feedthroughs. The Kryoflex™ line features a solder flange or weld flange in several variations and lengths, and can be soldered directly into aluminum packages, eliminating solder breakdown for mismatched feedthrough and chassis coefficients of expansion. **Kyle Technology Corp., Roseburg, OR. INFO/CARD #227.**

Frequency Counter Range Extender

The PM 9608B high frequency input option extends the frequency range of the PM 6660 series counter/timers from the standard 0.1 Hz to 120 MHz to 1.3 GHz. The PM 6660 series of frequency counters/timers is made up of three models: the PM 6669, PM 6665 and PM 6666. The PM 9608B is $325, and may be installed by the customer or factory. **John Fluke Mfg. Co., Inc., Everett, WA. INFO/CARD #226.**

High Performance Oscillator

The X01151 ovenized oscillator offers $\pm$ 3.0 x 10^{-8} stability over an extended temperature range of −55 to +95°C. The standard 10.0 MHz frequency can be modified to operate between 8 and 20



MHz. Maximum power consumption (steady state) is 5 watts at −55°C. **Piezo Technology, Inc., Orlando, FL. Please circle INFO/CARD #225.**

TO-5 Attenuator

EMC Technology introduces a series of DC to 2 GHz fixed attenuators in welded, hermetic TO-5 cases. Model CA3XXXX (XXXX = attenuation value) is available in attenuation values from 0.25 dB to 20 dB in 1 dB steps. Accuracy is $\pm$ 0.15, typical impedance is 50 ohms, and maximum VSWR is 1.30:1. Price is $14 in quantities of 100. **EMC Technology, Inc., Cherry Hill, NJ. Please circle INFO/CARD #224.**

Digitally Compensated Oscillator

This digitally compensated crystal oscillator is available for military and commercial applications at frequencies from 2 to 25 MHz. Worst-case stability is $\pm$ 5 X 10^{-7} from −40°C to +85°C, and power consumption is less than 60 mW. Price is $380 each (100s). **Thomson Electron Tubes and Devices Corp., Totowa, NJ. INFO/CARD #223.**

VSWR Meter

The CM-Z20 is a VSWR meter for measuring forward and reflected power and VSWR measurements for use with the CMT and CMTA radio-communication testers. VSWR measurements can be performed between transceiver and antenna at frequencies from 1 to 1000 MHz. The power range is 100 mW to 50 W with maximum resolution of 10 mW. The VSWR meter can be controlled by the IEC/IEEE bus or by the autorun control of the CMT/CMTA. **Rohde & Schwarz, Munich, West Germany. INFO/CARD #222.**

VHF Synthesizer

This low-phase-noise VHF synthesizer is available for frequencies from less than 90 MHz to greater than 154 MHz. The tuning range is limited to approximately 10 percent segments,

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RF Expo East

ENGINEER TO ENGINEER

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A Word About the Chairman

We are delighted to have Andrzej B. (Andy) Przedpelski as the Program Chairman for RF Expo East '89. He is Vice President, Development, at A.R.F. Products, Inc., at their Research and Development Lab in Boulder, Colorado.

Andy is familiar to many *RF Design* readers as the author of many tutorial articles on phase-locked loops, filters, matching networks, antennas and switches. He has been the magazine's Consulting Editor for over ten years and was the Program Chairman of the first ever RF Expo in 1985.

CONFERENCE TOPICS

Tuesday Morning

Filter Design
Modulators and Demodulators
IC Applications

Tuesday Afternoon

Power Amplifiers
CAD Methods
Frequency Synthesis

Wednesday Morning

Antenna Tutorial
Small-Signal Circuits
Receiver Design

Wednesday Afternoon

Transmission Systems
Phase-Locked Loops
Component Applications

Thursday Morning

Tuners and Combiners
Oscillators
Antennas

LOGISTICS

Headquarter Hotel:
RF Expo East's move to the TropWorld in Atlantic City means convenience and fun at reasonable rates for the attendees. Because the exhibits, meeting rooms, casino, restaurants and guest rooms are all under one roof, you'll find everything you need right at the hotel. Single or double rooms are $76.00/night.

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Atlantic City International Airport is served by US AIR from Pittsburgh; Piedmont Commuter from Baltimore/Washington Airport, Boston, MA, Albany, NY, and Burlington, VT; and Allegheny Commuter from Philadelphia International Airport. Bader Field Airport is served by Eastern Express from Washington National Airport, LaGuardia International, and Philadelphia International Airport; and Continental Express from Newark.

Travel Tips by Car:
From New York and North: Major interstate highways connect with the Garden State Parkway South to Exit 38 Atlantic City Expressway. Follow Expressway direct to Atlantic City.

From Philadelphia and West: Exit Pennsylvania Turnpike at Valley Forge Exit 24, onto Schuylkill Expressway (Rt. 76) through Philadelphia to the Walt Whitman Bridge, N/S Freeway to the Atlantic City Expressway and follow direct to Atlantic City.

Fundamentals of RF Circuit Design: Part I

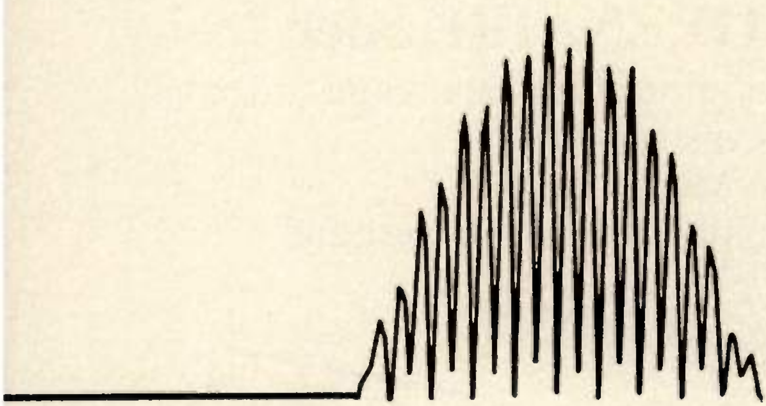
Course Outline:

Morning: **Introduction to High Frequencies**
Explains lumped and distributed components, parasitic effects, use of the Smith Chart, as well as theory and various physical realizations of transmission lines, with emphasis on practical design aspects. The session will conclude with a review of various components used in RF systems.

Afternoon: **Small Signal Amplifier Design**
From Smith Charts to computers, using S-Parameters, and covering unilateral and bilateral design using constant gain circles; computer-aided synthesis and optimization, stability considerations and more.

Instructor:
Les Besser, President
Besser Associates, Inc.

Date: Monday, October 23, 1989
Registration: 7:00 am - 10:00 am
Course Hours: 8:30 am - 5:00 pm



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Shuttles from Newark
For your convenience we will run one shuttle bus each day from the New Jersey Institute of Technology. Seats must be reserved in advance. For more information or to make your reservation call LeAnn Nowacki.

Fundamentals of RF Circuit Design: Part II

Course Outline:

Morning: A continuation of RF filter design, using structures beyond those introduced in the first course. Modern filter design, including computer-aided synthesis is also covered. Impedance transformation are extended to broad bands, using both lumped and distributed (transmission line) elements and complex terminations.

Afternoon: The small-signal amplifier design is also extended to broad bands; impedance matching and feed-back methods are discussed. Large signal amplifiers of Class-B and -C types are evaluated through the latest CAD tools. Other non-linear circuits, such as oscillators, mixers and frequency multipliers are introduced; their behaviors are examined both in the frequency and time domain.

Instructor:
Les Besser, President
Besser Associates, Inc.

Date: Tuesday, October 24, 1989
Registration: 7:00 a.m. - 10:00 am
Course Hours: 8:30 am - 5:00 pm

Computer-Aided Filter Design

Course Outline:

Introductory Topics:

- Lowpass prototype selection
- Transformation to highpass, bandpass & bandstop
- Group delay characteristics
- Effects and compensation of component unloaded Q
- Other component parasitics

Advanced Topics:

- Effects of transforms on filter amplitude and delay response
- Elliptic function filters
- Filter response symmetry
- Optimization
- Zig-zag minimum inductance bandpass filters
- Creating unique lowpass prototypes
- Multiple-section group delay equalizers

Instructor:
Randy Rhea
Circuit Busters

Date: Monday, October 23, 1989
Registration: 7:00 am - 10:00 am
Course Hours: 8:00 am - 5:00 pm

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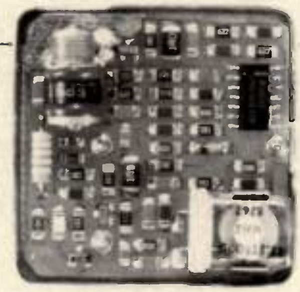

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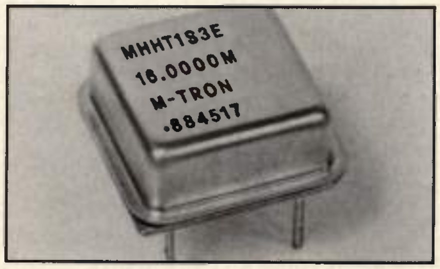
with step sizes from 2 kHz and switching speed from one end of the band to the other of less than 100 ms. Phase noise at 10 kHz offset is 132 dBc/Hz. **Communication Techniques, Inc., Whippany, NJ. INFO/CARD #221.**

High Dynamic Range MMIC Amplifier

MSA-1110 is a silicon bipolar MMIC amplifier which features a gain of 12 dB, third-order intercept point of +30 dBm and a typical noise figure of 3.5 dB at 500 MHz. It is intended for use in both 50 and 75 ohm microstrip systems. Price is $25 in 100-piece quantity. **Avantek, Inc., Milpitas, CA. INFO/CARD #220.**

Half-Size Oscillators

A new series of HCMOS oscillator



provides TTL compatible output at frequencies from 1.0 to 40 MHz. Supply voltage is 4.5-5.5 volts, and operating temperature is 0 to +70°C. **M-tron Industries, Yankton, SD. Please circle INFO/CARD #219.**

Class A Amplifier

PST introduces the Model AM1929-20 Class A amplifier that operates from 1000 to 2000 MHz. Power output is 20 watts at 1 dB compression, gain is 43 dB minimum, and the instantaneous bandwidth is 1000 MHz. Input/output VSWR is 2:1, gain flatness is ± 1 dB, and spurious signals are −60 dBc. **Power Systems Technology, Inc., Hauppague, NY. INFO/CARD #218.**

Helical Filters

Toko America introduces the adjustable 7HW double-tuned and 7HT triple-tuned helical filters that operate from 350 MHz to 1 GHz. They feature a top-adjustable, screw-type core for tuning, and a silver-plated brass cavity. Input impedance is 50 ohms. In quantities of 100, prices range from $4.00 for the 7HW to $5.75 for the 7HT. **Toko America, Inc., Mt. Prospect, IL. Please circle INFO/CARD #217.**

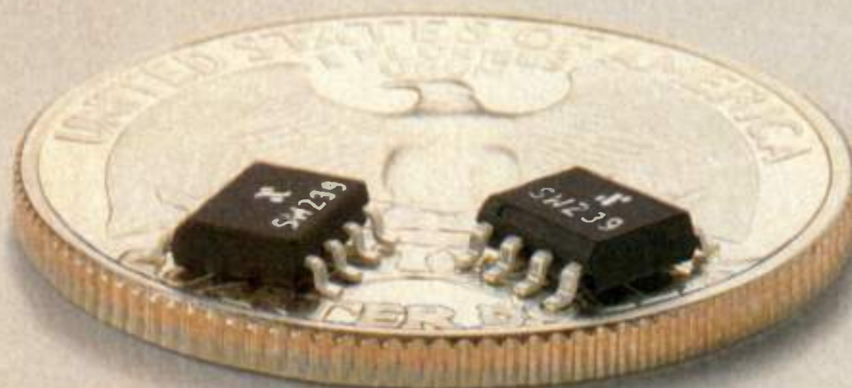
New TCVCXO and OCXO Lines

Temperature-compensated, voltage controlled oscillators and oven-controlled oscillators are offered for critical applications. The TCVCXOs are available for 1 to 150 MHz, with sine wave outputs. The OCXO line offers low aging and sine or logic outputs from 0.01 Hz to 30 MHz. **K & L Oscillatek, Olathe, KS. INFO/CARD #216.**

Coaxial Adapter

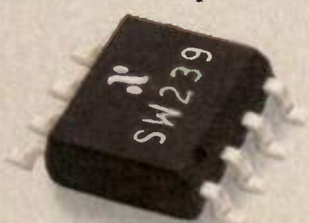
Pasternack introduces a Type SMC(P) to N(F) coaxial adapter that features low loss over DC to 8 GHz. Model PE9320

Two SPDT Bits.



ANZAC Does It Again!

ANZAC offers yet another GaAs Monolithic alternative to pin diode switch technology.



The SW-239 SPDT RF switch features nanosecond speed with only microwatts of DC power. Utilizing an 8-Lead SOIC package these units come ready for full automatic insertion.

MODEL SW-239 Performance Characteristics	
Frequency Range	DC-1000 MHz
Insertion Loss	0.5 dB
VSWR	1.1 : 1
Isolation	50 dB
1 dB Compression	+27 dB
Switching Speed	4 nsec TYP
Package	8 Lead SOIC (3.8mm x 4.8mm)

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has a brass nickel plated body, utilizes PTFE insulation and features a gold plated contact and a −65 to +165°C temperature range. In 100s, the adapter costs $23.96 each. **Pasternack Enterprises, Irvine, CA. INFO/CARD #215.**

Relay Package Prototyping Board

The Relay1 Quick Board is designed for the 8-pin, 0.4 in. X 0.8 in. standard industry relay package like the SRA-1 mixer. 50 ohm lines for the input and output can be used with either the PC mount or edge mount SMA, or BNC connectors. A socketed version is available. **RF Prototype Systems, San Diego, CA. INFO/CARD #214.**

25-80 MHz VCXOs

The HV50 series (HCMOS) and the EV535-100 series (ECL) cover the frequency range of 25-80 MHz. The HV50 is available with center frequency stability to ±5 ppm over 0 to 70°C. In 10-piece quantities, prices range from $65 to $125, depending on frequency. **Connor-Winfield Corp., Aurora, IL. Please circle INFO/CARD #213.**

SAW Bandpass Filters

Sawtek introduces a line of 70 MHz filters which features more than 40 bandpass filters with 3 dB bandwidths ranging from 0.25 to 40 MHz. Unit prices are under $100 in quantities of two or more. **Sawtek, Inc., Orlando, FL. Please circle INFO/CARD #212.**

Crystal Clock Oscillators

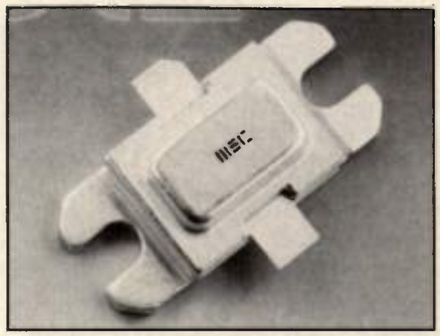
NEL unveils CMOS-compatible crystal clock oscillators that range in frequency from 244 Hz to 3.5 MHz. Model HS-400 is available for tolerances from ± 0.005 percent. The line is packaged in welded metal case. **NEL Frequency Controls, Inc., Burlington, WI. Please circle INFO/CARD #211.**

Mechanical Phase Shifter

The Model FPS4387 mechanical phase shifter features a DC to 3.5 GHz operating range with maximum insertion loss of 0.9 dB and VSWR of 1.5:1. The minimum available phase shift is 288 degrees/GHz, equivalent to 0.8 nanosecond delay variation. An SPDT switch indicates the travel limit. **Sage Laboratories, Inc., Natick, MA. Please circle INFO/CARD #210.**

L-Band Power Transistor

The MSC 1214-175 is a high-power refractory/gold metallized Class C transistor designed for pulsed output and



driver applications in L-Band radar. It is capable of delivering 180 watts of output power over the 1215 to 1400 MHz range. This matched device has a typical gain of 7.8 dB at a V_{cc} of +40 V and a typical collector efficiency of 50 percent. **Microwave Semiconductor Corp., Somerset, NJ. INFO/CARD #209.**

Miniature Filters

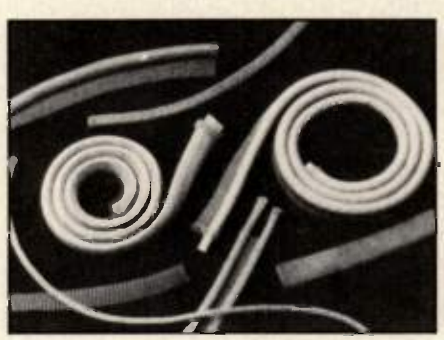
Wavetek RF Products introduces a new version of the Ultramin® line of miniature filters aimed at telecommunications applications. The commercial Model B2 bandpass filter package is available with a center frequency range of 1 to 500 MHz, 5 to 40 percent bandwidth and ultimate rejection of 50 dB absolute. In this standard range, costs are as low as $30 each. **Wavetek RF Products, Inc., Indianapolis, IN. INFO/CARD #208.**

Temperature-Compensated Crystal Oscillator

This TCXO, the TO-2100-1, is available in frequencies from 6 to 24 MHz. Stability as good as ± 1.0 ppm is available over the −40 to +85°C range. Output voltage is 1.8 $V_{p\text{-}p}$ with a 1 kohm load. The oscillator occupies 0.28 cu. in. and draws 35 mW power. **Murata Erie North America, Inc., State College, PA. INFO/CARD #207.**

EMI/RFI Shielding

Multi-Point® VII is a gasketing material that provides EMI/RFI shielding protec-



tion for metallic and non-metallic enclosures. It is produced in two constructions: as an all-metal gasketing consisting of knitted wire mesh strips, and as a composite knitted metal mesh and elastomer gasketing consisting of two layers of knitted mesh enclosing a core of neoprene sponge or silicon sponge. **Conductive Systems, Inc., Amsterdam, NY. INFO/CARD #206.**

806-941 MHz Cylindrical Antenna

Model BCD 80612 is an omnidirectional antenna with 10 to 12 dB gain over the entire 806 to 941 MHz range, for cellular, conventional, and trunking. Standard antenna input power is 250 watts with type N connectors. 7/16 or EIA 7/8 in. connectors allow operation at 1000 watts or 2000 watts, respectively. In a typical trunking system, the antenna can handle up to 40 channels with a 2000 watt input. **Antel Inc., Crofton, MD. INFO/CARD #205.**

Frequency Synthesizer

Syntest introduces the SM-102 frequency synthesizer, providing a TTL square wave signal into a 50 ohm load over the 0.1 Hz to 16 MHz eight-decade ranges. Programming is achieved by TTL compatible, parallel BCD lines. Maximum spurious is −60 dBc and settling time is 10 ms to within 10 percent of a frequency step. Single unit price is $528. **Syntest Corp., Marlboro, MA. INFO/CARD #204.**

SC-Cut Crystal Oscillator

The Model 2880063 oscillator incorporates SC-cut crystal technology in a low-noise ovenized circuit. Frequency range is 5 to 10 MHz and typical phase noise for 5 MHz is −128 dBc/Hz at 10 Hz. Typical vibration sensitivity is 2 x 10^{-9}/g at 10 MHz. In small quantities, price ranges from $450 to $650. **Piezo Crystal Company, Carlisle, PA. Please circle INFO/CARD #203.**

rf software

Smith Chart Program

MicroSmith is a Smith Chart program written for the IBM PC/XT/AT and compatibles with DOS 2.0 or later. The program is a working Smith Chart optimized for impedance matching applications. Component values and frequency may be tuned from the keyboard. The chart shows both the graphic output and numerical values for complex impedances or admittances, reflection coefficient, return loss and VSWR. Circuit editing generates an on-screen schematic diagram. Circuit components supported are R, L, C, frequency-independent reactances, transmission lines including stubs, and a dipole antenna model. The software is priced at $29. **Hayward Electronic Systems, Inc., Beaverton, OR. INFO/CARD #199.**

New Version of Filter Program

Circuit Busters introduces Version 2.1 of =Filter=, a filter synthesis program. The new features of this program include automatic design of group delay equalizers for computed or measured filter data, Blinchikoff 30 to 70 percent bandwidth constant delay bandpass filters, singly terminated Cauer-Chebyshev filters, and computation of filter effective noise bandwidth. Version 2.1 is priced at $495 and owners of Version 1.2 can upgrade at a special price. **Circuit Busters, Inc., Stone Mountain, GA. INFO/CARD #198.**

Communication Simulator

The Workstation Communication Simulator (WCS) supports the interactive design and evaluation of complex digital communication systems. WCS provides the user with the ability to model and analyze general point-to-point communication links for military, commercial and space applications. The system allows high-level graphical access to a variety of functional modules to implement the various functions in an end-to-end digital communications link. The software simulates the overall operation of the link design. WCS runs on an IBM PC/XT/AT or compatible with EGA and math coprocessor. It costs $4500. **ICUCOM Corp., Wynantskill, NY. Please circle INFO/CARD #197.**

rf literature

EMI/RFI Problems and Solutions Video

The basic concepts of EMI/RFI problems in electronic cabinets, along with industry solutions, are the subject of this VHS video tape from Equipto. Requirements for meeting FCC and military EMI/RFI specifications are highlighted. EMP and TEMPEST concerns are also covered, with discussions on alternatives for solving these problems. An EMI/RFI technical guide which features shielding, galvanic compatibility, I/O options, product specifications and shielding test results is included. **Equipto Electronics Corp., Aurora, IL. Please circle INFO/CARD #196.**

Spectrum Analyzer Booklet

The Tektronix Spectrum Analysis Primer provides a basic tutorial on spectrum analyzers. A description of the instrument, together with a glossary, is included. Other sections featured include spectrum analyzer amplitude calibration factors, a dB conversion table, impedance conversions, and a frequency allocations table. Tutorials on amplitude modulation, frequency modulation, pulsed RF, and intermodulation distortion intercept points are also included. **Tektronix, Inc., Beaverton, OR. INFO/CARD #195.**

Chip Inductors Bulletin

This bulletin from Sprague-Goodman features their line of Surfcoil^{R} chip inductors. Bulletin SG-800A provides features, specifications, inductance and Q tables, and schematic drawings. In addition, a page is devoted to carrier and reel specifications. **Sprague-Goodman Electronics, Inc., Garden City Park, NY. INFO/CARD #194.**

RF Filters Catalog

RF/88 is a catalog that describes a variety of filters for aerospace, defense electronics and VHF/UHF radio. Medium- and high-power lowpass filters (up to 1400 watts) for suppression of transmitter harmonics are featured. Diplexers for combining two transmitters or receivers to a common antenna or two antenna ports to a cable for remote transport are also described. **Microwave Filter Company, Inc., East Syracuse, NY. Please circle Please circle INFO/CARD #193.**

Electronics Buying Guide

The 1989 Newark catalog contains more than 1,100 pages of dimensions, specifications and descriptions on over 100,000 products from more than 240 manufacturers. It includes 7,900 new products and 13 new product lines for 1989. **Newark Electronics, Chicago, IL. INFO/CARD #192.**

Active Probe Data Sheet

This data sheet from Marconi describes a 1 GHz active high-impedance probe for IF, HF, VHF and UHF applications. Specifications together with a typical harmonic distortion and intermodulation performance curve plus accessories and ordering information are included. **Marconi Instruments, Allendale, NJ. INFO/CARD #191.**

Transistor Data Sheet

MSC introduces a data sheet outlining the performance of the MSC 1517-35 device. This Class C transistor is designed for high power CW telemetry/telecommunications applications. **Microwave Semiconductor Corp., Somerset, NJ. INFO/CARD #190.**

Measurement Accessories Catalog

This edition of HP's RF, microwave and MM-wave catalog (#5953-2346) features more than 500 products that range in frequency from DC to 110 GHz. New products featured include an SPDT coaxial switch, 11, 70 and 90 dB attenuators for 40 GHz, and a family of planar-doped-barrier coaxial detectors. Selection guides and product sections include fixed and step attenuators, adapters, detectors, power sensors, probes, and 75 ohm components. **Hewlett-Packard Company, Palo Alto, CA. Please circle INFO/CARD #189.**

Signal Processing Components Databook

Comlinear introduces their 1989 databook which contains specifications on operational amplifiers, buffer amplifiers, linear amplifiers, track and hold amplifiers, A/D converters, D/A converters, and encased amplifiers. A product selection guide is included. Also featured are various related application notes. **Comlinear Corp., Fort Collins, CO. INFO/CARD #188.**

Broadcast Catalog

Richardson Electronics has published its new broadcast catalog. The company distributes broadcast products for companies such as Acrian, Burle, Hitachi, Jennings, Motorola, MPD (G.E.), Panasonic, SGS-Thomson, Siemens and Varian. **Richardson Electronics, Ltd., LaFox, IL. INFO/CARD #187.**

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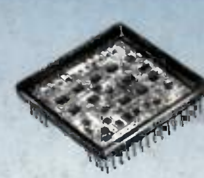

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