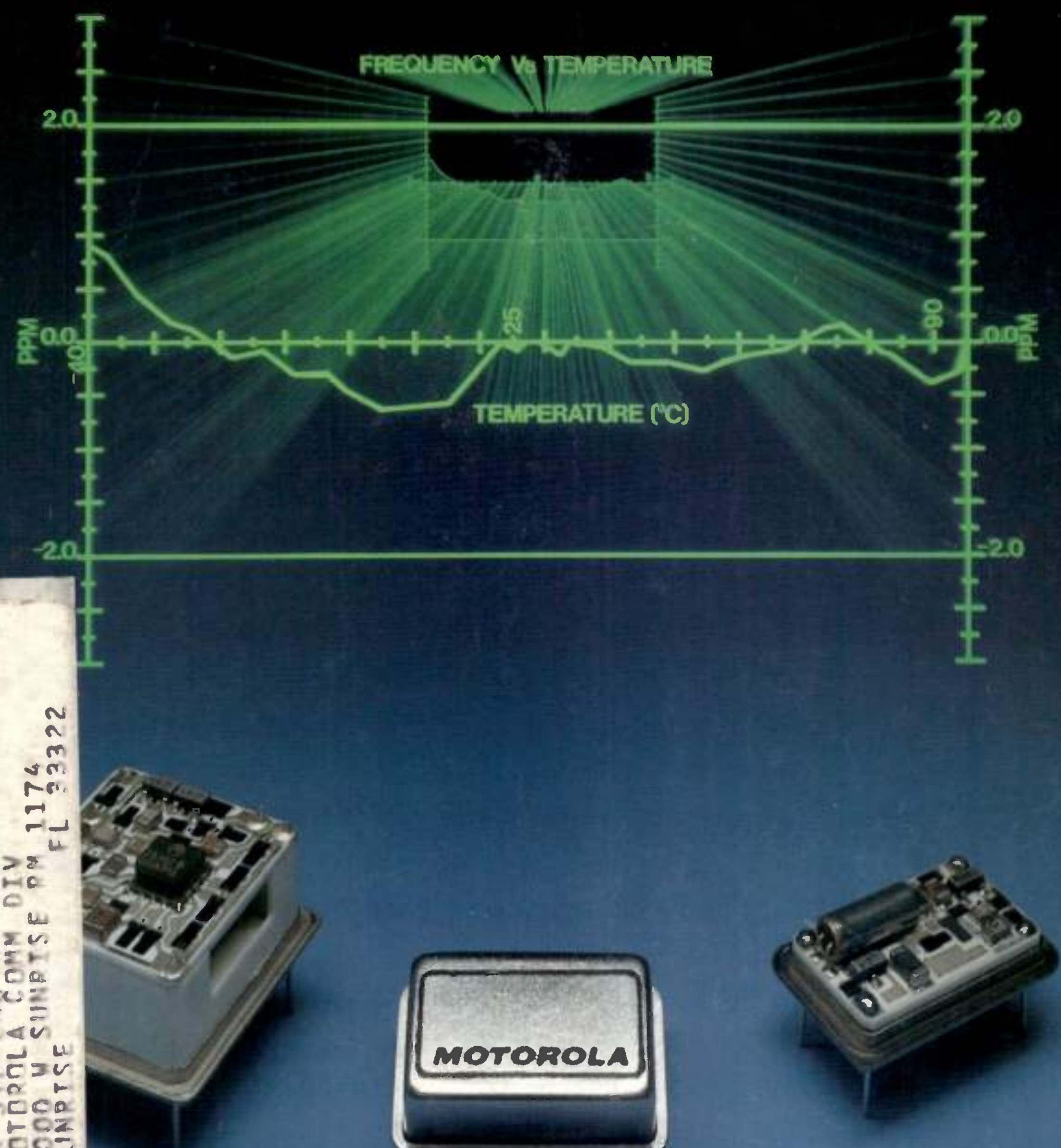
rf design

ideas for engineers

March 1985



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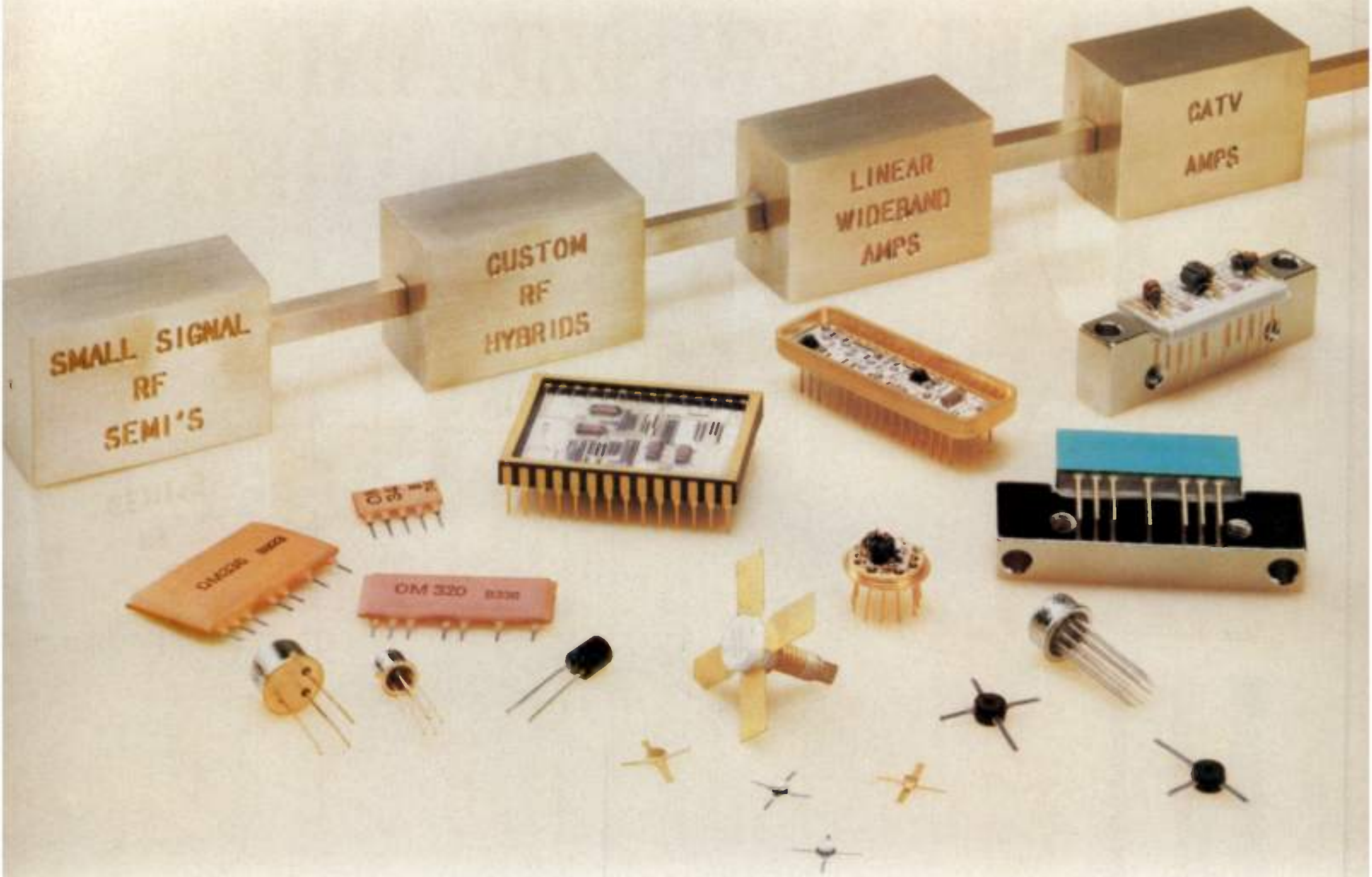

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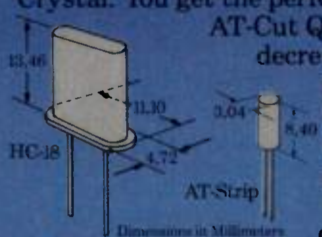

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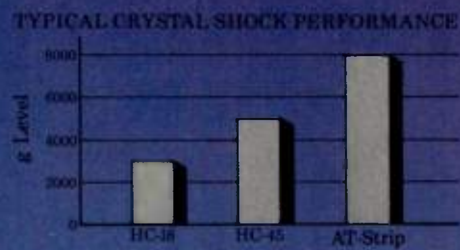

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INFO/CARD 2

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rf design

March 1985

Scenes from RF Technology Expo 85 — Anaheim, Calif.

Cover

This month's cover features Motorola's MDO-2 (left) and the new MDO-3 (right) temperature-compensated crystal oscillators. Motorola considers these MDO products to be the best in TCXO performance in their cost range. The MDO-3 is described in this month's Special Report, along with other recently announced oscillator products. Many of the considerations mentioned in Dennis Marvin's article on frequency-temperature performance determination in crystal oscillators were incorporated in the MDO-2 series. (Photo courtesy of Motorola, Inc.)

Features

Departments

R.F. DESIGN (ISSN: 0163-321X USPS: 453-490) is published monthly plus one extra issue in June. March 1985, Volume 8, No. 3. Copyright 1985 by Cardiff Publishing Company, a subsidiary of Argus Press Holdings, Inc., 6530 S. Yosemite Street, Englewood, CO 80111 (303) 694-1522. Contents may not be reproduced in any form without written permission. Second-Class Postage paid at Englewood, CO and at additional mailing offices. Subscription office: 1 East First Street, Duluth, MN 55802, (1-800-346-0085). Subscriptions are sent free to qualified individuals responsible for the design and development of communications equipment. Other subscriptions are: $15 per year in the United States; $25 per year in Canada and Mexico; $25 per year for foreign countries. Additional cost for first class mailing. Payment must be made in U.S. funds and accompany request. If available, single copies and back issues are $5.00 each (in the U.S.). This publication is available on microfilm/fiche from University Microfilms International, 300 N. Zeeb Road, Ann Arbor, MI 48106 USA (313) 761-4700.
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rf editorial

Where Do You Learn About RF Circuits?

James MacDonald
Editor

The biggest surprise at the RF TECHNOLOGY EXPO was the response to the Fundamentals of RF Design course. Based on pre-registrations, the room was set up for a maximum of 350. By starting time the first day the room was filled and more than 100 conference attendees were in the hall hoping to register. We were able to add some chairs, but many who had not pre-registered had to be turned away.

The situation was similar when the course was repeated the next day. By then the room had been expanded to seat 500, anticipating the overflow from the first day. All seats were filled and remained filled throughout the day.

On one hand, this overwhelming response reflected the response to the EXPO as a whole. Attendance exceeded our expectations, if not our hopes. On the other hand, it provided dramatic proof of a serious lack of educational opportunities for RF engineers. The audience composition ranged from young recent graduates to university professors with well-established reputations in electrical engineering.

As I talked with attendees the same fact was mentioned over and over; electrical engineering students have little opportunity or incentive to study RF circuit principles. In many universities the courses simply are not available. Where they are available advisors do not encourage such courses.

Current employment opportunities apparently are greater and salaries higher in the digital and power fields, but this not-very-widely promoted course in RF design fundamentals attracted more than 800 engineers and others involved in designing RF equipment, some from foreign countries. Clearly there is a significant lack of educational opportunities in RF circuit theory and design. We think engineering colleges should take a look at this educational need.

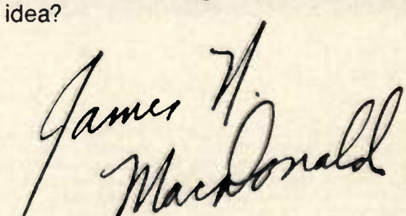
Proceedings of the sessions, including the Fundamentals of RF Design course, will be published soon. RF designers who could not attend the EXPO may want to order a copy. Each session was tape recorded, as well. Tapes may be ordered from Meyer Communications Corp., 13791 E. Rice Place, Aurora, CO 80015.

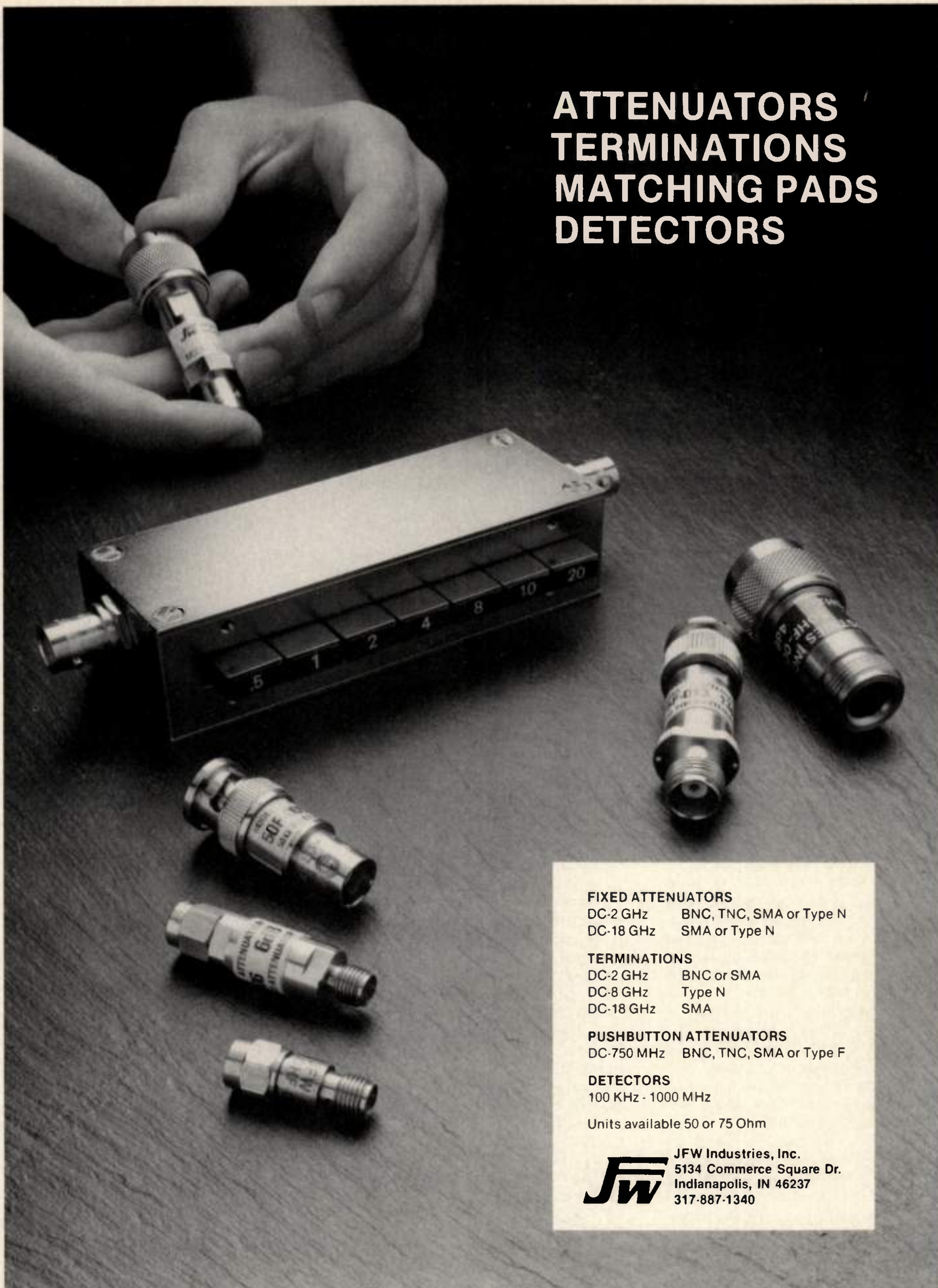
During World War II many amateur radio operators were hired by defense contractors to build communications gear. Even fewer engineers were trained in RF principles at that time. Amateurs knew how to build equipment, even if they did not thoroughly understand how it worked. Many of them eventually became engineers.

This is one reason so many RF designers are amateur radio enthusiasts. Most of those who worked during WWII are retired, but they passed their enthusiasm on to the younger generation. During EXPO 85 quite a few hams were seen with handheld 2-meter gear and others were heard on the band with mobile and base station rigs.

Carl Lodstrom, SM6MOM/W6, has suggested that we select a frequency for simplex operation during the next EXPO and publicize it. Carl said such a frequency seemed to be chosen by chance at EXPO 85, perhaps because people heard others talking on it and joined the net. He said it was a great way for hams to get together for dinner or meetings.

What do *RF Design* readers think of this idea?

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INFO/CARD 5

rf publisher's note

That Old Black Magic — Losing Its Old Spell

Keith Aldrich
Publisher

Something happened at the RF TECHNOLOGY EXPO 85 in January: something which spells the end for that old myth about RF technology being a "black art" which practitioners exercise by instinct, or by "the seat of the pants."

Call it a new generation. Call it the light of rational analysis breaking like a dawn to dispel the blackness. Whatever you call it, it became obvious at the EXPO that "tweaking" and "tuning" approaches to RF design are giving way to analytic approaches, in which the vagaries of RF propagation are not regarded as unexplainable mysteries, but phenomena which can be characterized, predicted, and harnessed with precision. The end result is designs which take days instead of weeks and which are proved in the lab instead of the field.

The new generation factor was evident in viewing the 800-plus attendees who listened with such close attention to Les Besser and other instructors at the "RF Fundamentals" course. While there was a liberal sprinkling of gray heads . . . more, perhaps, than the title of the course would lead you to expect . . . the studiously attentive faces were mainly young, young, young. Not college young. The badges read "Hughes" and "Rockwell" and "General Dynamics," and the exhibitors who encountered them later in their booths exclaimed over their knowledgeable questions and their purchasing power (one instrument manufacturer announced sales of $250,000 worth of equipment the first day). No, these were real engineers . . . just young — and *eager to learn.* Les Besser was up there talking about constant gain circles, and S-parameters, and computer-aided design, and they were eating it up.

Just a couple of years ago, RF engineers were characterized as being hostile to such innovations, on the average, preferring to "do it their way." And, on the average, they have successfully resisted the innovations. Instructor Besser asked for a show of hands, toward the end of his lecture, to see how many attendees commonly applied tolerance analysis to check out their designs. Only a small fraction raised their hands. "About three percent," Les computed; "about the national average." He went on to preach the gospel: how his listeners could cut down their failures in the field by applying some of the methods he was espousing.

It had the ring of one of those futile lectures about littering or something. Except for one thing. These young people were listening carefully, not yawning and fidgeting. And you realized, if you thought about it, that these people were not *inherently* antagonistic to computers, and modeling and simulation. They grew up with computers. They are already competent EEs, for the most part. It is only RF that is new to them. They are listening so carefully to Les because they are going to *practice* what he is preaching.

That is good news for their employers, and for the RF boom that was so apparent in the mood of RF TECHNOLOGY EXPO 85. While much of the electronics industry is faltering, and while digital engineers in many parts of the country are being thrown out of work, manufacturers of RF equipment can hardly keep up with the demand. To keep up with what is needed from them RF engineers must be as productive as modern technology can make them. We are pleased to have been of service in helping bring them up to speed.

Keith Aldrich

Publisher
Keith Aldrich

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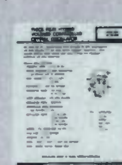

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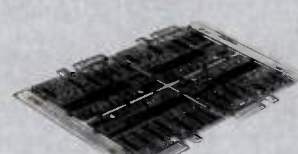

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rf letters

Editor:

I enjoyed the "Frequency Synthesized Signal Generators" article in the November/December 1984 issue of *RF Design.*

I was dismayed, however, to find that Eaton Electronic Instrumentation Division's signal generator products were not included in the story, or the product briefs which followed, although I responded to your request for information.

I am sure the oversight was unintentional but, in the event you are planning future coverage of this product line, I have included data sheets descriptive of both the Model 460 AM/FM ΦM Signal Generator and the 360, 380 Series of Direct Synthesized Signal Generators for your future reference. The 380 Series is particularly exciting offering 20 microsecond switching to 4 GHz.

If I can ever provide additional information on Eaton's broad line of test and measurement instrumention, please call me without hesitation.

Sincerely,
Patricia W. Ullrich
Marketing Communications Manager

Because of the large response from manufacturers to our call for new product information we must set limiting guidelines for our special reports. We attempt to keep our readers informed of the most recent product developments. A number of products were not included in the special report on "Frequency Synthesized Signal Generators" because they had been on the market long enough for most RF engineers to know of them. We welcome comments from manufacturers and readers on this subject. Readers wanting information about the products mentioned in Patricia Ullrich's letter may circle info/card #177. — editor

Sweet Smell of Success at RF Technology Expo 85

Even the night before it opened, at a crowded VIP reception in one of the Disneyland Hotel's posh ballrooms, exhibitors and speakers were congratulating executives of sponsoring Cardiff Publishing Company, *RF Design* magazine staff members and each other. Cardiff president Robert A. Searle, veteran of many such shows, said late that night, "They're declaring it a success before it happens. When they do that it's a good sign."

The stage was set. In the exhibit hall 60 exhibits were poised in 75 booth spaces. Elsewhere in the complex five large rooms were set up for the concurrent technical sessions accommodating the 71 papers to be given at the first RF TECHNOLOGY EXPO.

Shortly after dawn the next day, long before registration desks opened, the line of attendees began to form. The hundreds who already had registered by mail were there just to pick up their badge and program, but they were joined by hundreds starting from scratch.

The final attendance figure was 1,300, plus 400 exhibit personnel and speakers, and as Convention Manager Kathy Kriner said, "Nobody expected them show up all at once!"

The opening day jàm-up was alleviated by dispatching those who were there for the Fundamentals of RF Design course directly to the classroom *sans* admission, to return for registration at the noon break.

In the hotel's huge Embassy Room, classroom for the fundamentals course, instructor Les Besser was staggered as an overflow crowd topping 400 poured into a space set up for 350. Adapting admirably, the president of Microwave Educational Programs held the jammed-in crowd spellbound through a four-hour lecture on Small Signal Amplifier Design.

In mute testimony to the value of the material they were receiving, the crowded participants sat silently and seemingly motionless as instructor Besser revealed the mysteries of what others have called the "Black Art." The art became a science.

In other rooms, some almost as crowded, three papers were being given concurrently, truly offering something for everyone. Attendees could go from room to room, from topic to topic, picking up the particular knowledge needed for the job back home. Four papers were delivered in each room.

Registration continued through the morning, as exhibitors prepared for the noon opening. Many of them later said they were not prepared for the number and caliber of people who crowded their booth constantly until the exhibits closed at 6:00 p.m.

"Every person I talked to was a qualified customer," was a common remark from exhibitors.

Many discovered they had brought too few people to talk to the visitors and too little printed material to hand out. Frantic calls went out for help and literature. Business cards became unexpectedly scarce items.

While attendees enjoyed lunch at the variety of famous restaurants in the hotel complex, hotel employees expanded the Embassy Room to provide comfortable seating for the extra 100 in the Fundamentals class. University of Colorado Professor K.C. Gupta started the afternoon session with a detailed explanation of striplines and microstriplines. He was followed by Joe Johnson, president of Microwave Modules and Devices, explaining high power amplifier design: John Morton, engineering manager for Microsonics, Inc., who gave an overview of oscillators; and Carl Erickson Jr., with a complete description of surface acoustic wave devices.

From the silence that prevailed all day the intensely concentrating crowd of 400 could have been a group of six. Very few had left by the end of the session, and those who left may have gone to catch a particularly interesting presentation in one of the other sessions.

When all papers had been presented most attendees gathered at the exhibit hall for free refreshments and relaxed conversation. Even now the exhibit booths were crowded, although the pace was slower. Some exhibitors had made thousands of dollars worth of unexpected sales that afternoon. They were pleasantly tired, not just from the constant flow of visitors to their booths but also from the depth of the questions they had to answer.

"You were smart to only open the exhibits from noon to 6:00," one exhibitor told *RF Design* publisher Keith Aldrich. "I don't think we could have taken this all day."

Attendance at the RF Fundamentals course was even larger the second day. Once again the room was expanded as

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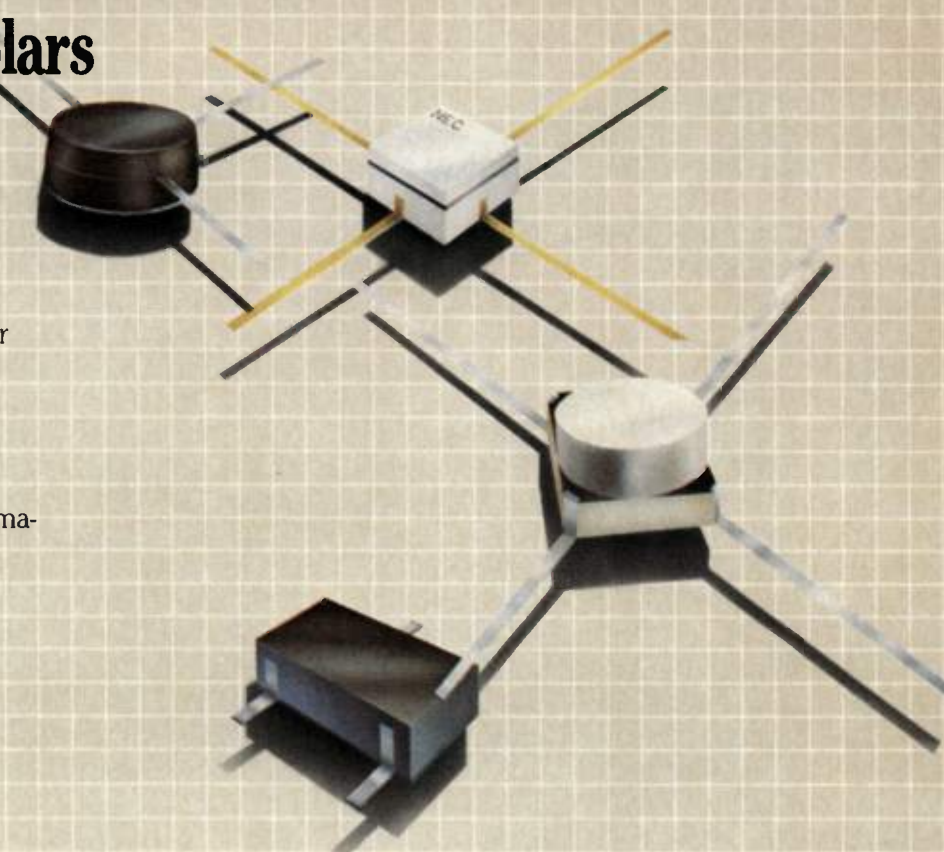

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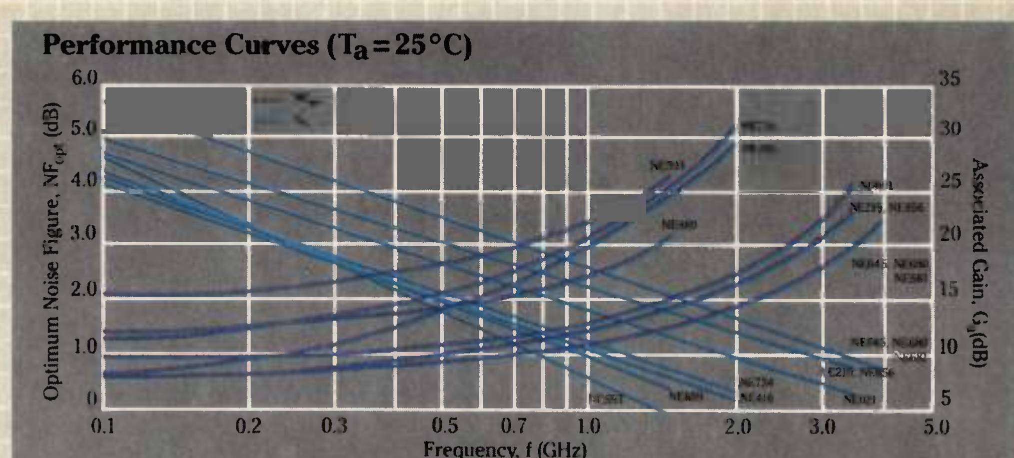

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INFO/CARD 13

hotel employees removed walls and brought in chairs. When they had finished, nearly 500 chairs were occupied.

It was remarkable that in spite of the last minute rearrangement of the room, leaving about 100 without tables to write on, there were no complaints. Those who were there considered themselves lucky to receive the information, regardless of the circumstances.

As many as 200 attended some of the other sessions, covering the general topics of circuits, transmitters, computer-aided design, antennas, techniques and components. Sessions were arranged both days so conference attendees could go from room to room and hear papers on each topic. RF TECHNOLOGY EXPO registration badges could be seen everywhere in the hotel complex as attendees went from session to session or to the exhibit hall.

His work was over, but program chairman Andy Przedpelski could be seen many places during the EXPO, going to sessions or touring the exhibits or just joining one of the many spontaneous discussion groups that seemed to develop wherever attendees came together. He had put together a valuable program, and now he could enjoy and benefit from it along with those he had done it for — the RF designers who came from as far away as China and Israel to attend.

It was no surprise to the staff that 360 copies of the proceedings had been ordered by those in attendance, or that 67 percent of the exhibitors said they definitely would be back for EXPO 86. Another 30 percent said they probably would be back. One who said he probably would not return said his products were "too microwave" for the attending engineers. He said most of those he talked to worked in frequencies "from 2 GHz down."

Of course, that was the idea of the RF TECHNOLOGY EXPO. It was the first show held specifically for engineers working from 2 GHz down. But Cardiff officials say it won't be the last one.

Keith Aldrich said, as the show was closing Friday morning, "This phenomenon confirms the premise of the show — that there is a desperate need for RF engineering information that is not being met, among engineers who have been pressed into service without adequate training in RF."

Dates and site of the RF TECHNOLOGY EXPO 86 will be announced in next month's issue of *RF Design*.

Dow-Key Becomes Subsidiary of Kilovac

Dow-Key Microwave Corp. was incorporated Jan. 2, as a wholly owned subsidiary of Kilovac Corp., Carpenteria, Calif. Sales manager Jack Dysart announced the company's new status at a news conference at RF TECHNOLOGY EXPO 85. The company previously had been a division of Kilovac involved in the design and manufacture of coaxial relays and switches for the RF and microwave communications, radar, electronic warfare and test equipment markets. The company recently moved into a large facility in Carpenteria with class 100,000 clean room facilities and MIC capability.

Dysart said the company is introducing the concept of intelligent relays which integrate higher levels of digital, analog and microwave integrated circuits into their existing line of electromechanical switches designed to meet Mil-S-3928.

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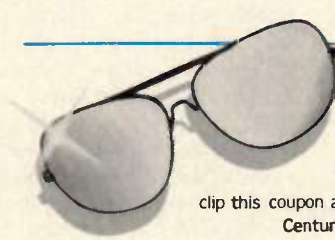

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INFO/CARD 15

New Waves In Oscillator Design

By James N. MacDonald

It starts with an amplifying device, usually a semiconductor. Part of the current is tapped from the amplifier output and fed back into the input. This feedback current is passed through a circuit designed to be resonant at a specific frequency. The resonant frequency within the band of frequencies in that current is again boosted by the amplifier and the process is repeated. If the amplifier provides sufficient gain and the current remains in phase, the circuit oscillates continuously at that frequency. This is the common oscillator circuit.

There are three types of oscillator circuits in general use by RF engineers, and nearly all receiving and transmitting systems as well as many other devices use one or more of each type. The types are based on inductor-capacitor (LC) circuits, crystals and surface acoustic wave resonators (SAWs). All types are widely used and new designs are being developed regularly from each type.

When first developed each type of circuit represented a new wave in oscillator design. The next new wave, described later in this report, will probably combine microprocessors with analog oscillator circuits. Between these major developments improvements have tended to involve better manufacturing methods, resulting in smaller and more stable devices to meet increasingly rigorous military and commercial requirements. Engineering design improvements have made the better manufacturing methods possible.

This report will discuss some of the latest developments in oscillators designed to operate primarily in the lower RF frequencies (through UHF). The distinction is difficult to make since design and manufacturing improvements are steadily extending the operating ranges of crystal and SAW oscillators. In general, we will not discuss atomic standards and oscillators designed to operate primarily above 2 GHz.

LC Circuits

The LC circuit is the first oscillator circuit learned by an RF designer. A coil and a capacitor connected in an AC circuit will resonate at a certain frequency determined by the reactance of each component. The resonant frequency is reinforced by the coil and capacitor, while other frequencies meet impedance. It is a simple circuit and relatively easy to design. The many improvements that have been made in this type of oscillator have depended mostly on the creativity of circuit designers. The LC circuit, sometimes called the tank circuit, can be used at any frequency, but in and above the UHF frequency range the electrical characteristics of inductors and capacitors begin to change and circuit design becomes more complex. At higher frequencies discrete components are replaced by thin-film deposited on small circuit boards or by cavity resonators.

The LC circuit with discrete coils and capacitors often is the only practical oscillator for low-priced, low frequency equipment where size is not a major consideration. A problem with this circuit is frequency instability in environments of wide temperature variation and physical shock. However, the LC circuit is still the choice of many RF designers in applications where these disadvantages are not significant or can be compensated for.

Quartz Crystals

For many years the quartz crystal oscillator has been the workhorse of the industry. A piece of quartz crystal will vibrate when AC voltage is applied across it. The frequency of vibration depends on the dimensions of the crystal piece, a slice from a naturally formed crystal. Current alternating at the natural vibrating frequency of the crystal will be in resonance with the crystal. The crystal functions like an LC circuit in a single element.

Held between metal electrodes the quartz crystal acts like a capacitor. It has very high Q and, consequently, good frequency selectivity.

Quartz crystal oscillators are also highly stable, holding on frequency under physical shock that might change the resonating frequency of an LC circuit. However, the resonant frequency of a crystal changes as its temperature changes. Since electronic equipment often is subjected to wide temperature variations, this tendency must be compensated for.

Frequency-temperature changes can be reduced by controlling the oscillator temperature, varying oscillator tuning voltage or using a phase lock loop. The PLL has been discussed in several previous articles, so it will not be discussed here. Oscillator temperature can be controlled by containing the oscillator in an "oven," a metal chamber kept at a constant temperature. Tuning voltage can be varied by a thermistor network reacting to oscillator temperature changes.

This issue's cover features Motorola's new line of temperature-controlled crystal oscillators. As the latest addition to their line of hermetically sealed oscillators, the MDO-3 offers ±5 ppm frequency stability over a –30 to +70 ° temperature range, company officials say. This stability is said to be inclusive of variations in temperature, voltage, load and humidity. Measuring 0.82 in. x 0.52 in. x 0.32 in. with DIP compatible pin out, this oscillator is said to be ideal for those applications in which low cost and small size are required. Motorola claims aging rates for this oscillator of ±1 pp.m

The MDO-3 operates from a 5 volt power supply over a frequency range of 10 to 20 MHz. It uses the AT-Strip crystal to achieve reductions in size and improvements in mechanical shock survivability over previously offered designs. The all-metal welded package provides a full hermetic seal which minimizes RF radiation while keeping the oscillator on frequency in environmental extremes of temperature and humidity.

The MDO-3 can be adjusted through the use of an external potentiometer, or it can be electronically tuned, providing AFC capability. Motorola says the automated manufacturing, proprietary thick film hybrid design and use of the AT-Strip quartz crystals have resulted in reduced parts count for cost savings with improved field reliability.

These are the kinds of improvements crystal oscillator manufacturers are making.

Often the designer wants the oscillator circuit frequency to be variable. This can be done by changing the capacitance of the circuit, either mechanically, with a variable capacitor, or electronically, with a varactor.

At the higher frequencies the crystal is so small and thin it is subject to breakage from its own vibrations. To avoid this problem crystal manufacturers obtain higher frequencies from a harmonic overtone of a crystal that resonates at a lower frequency. A frequency multiplier is used to select and amplify this harmonic overtone.

An analysis of new products recently announced shows that crystal oscillator manufacturers generally are striving for high stability, low power consumption, small size and fast warm-up.

Greenray Industries' new digitally compensated crystal oscillator is said to offer frequency-temperature stabilities approaching those of ovenized oscillators while consuming only a small fraction of the input power required to operate an oven. The advantage, Greenray says, is that the oscillator is on frequency instantly, eliminating the wait for an oven to stabilize.

The angle at which a slice is cut from a crystal is one of the factors determining its frequency-temperature stability, and many different cuts are used. CTS Corp., Knights Division, says their relatively new doubly-rotated S-C cut crystal resonator provides fast warm-up and excellent temperature and long-term stability in an ovenized oscillator. They say their TCXO design can consistently be produced to less than $\pm 4 \times 10^{-7}$ over a –55 to +85° temperature range for the life of the oscillator, even after the center frequency has been adjusted to account for aging.

As a custom manufacturer, CTS/Knights reports receiving more than a few inquiries for such devices as:

A small, low power, fast warm-up ovenized oscillator that can withstand the severe environments encountered in tactical jet fighters;

An oscillator packaged and mounted to permit mechanical shock operation of 3000 Gs;

A synthesized multiple-output device using monolithic VCOs to lock to a stable ovenized reference;

A leadless chip-carrier crystal oscillator — a complete quartz crystal oscillator in a ceramic chip carrier package.

CTS/Knights considers the requirements described above to be among the factors determining the future of the quartz crystal oscillator market.

Seiko Instruments has developed a series of oscillators using a GT-cut crystal, which company officials say provides good frequency-temperature characteristics. Seiko says it has developed a photolithographic fabrication process that permits the high-accuracy dimensioning required to mass produce the GT-cut.

Wenzel Associates has developed a small, low power ovenized oscillator they call the Small Fry. Officials say the Small Fry uses a 3rd overtone crystal which provides aging of 5×10^{-10} per day, with close in phase noise that can be specified to –135 dB at 10 Hz for 10 MHz oscillators. They attribute the performance to a tiny vacuum flask that insulates the oven and provides a five-fold improvement in thermal conductance over conventional foam and to a low flicker crystal exciter design.

The Vari-L Company has just announced a new line of voltage controlled oscillators covering frequencies as low as 25-50 MHz. The company says the low power dissipation, less than 200 mW, helps them meet rigid military specifications of –55 to +100°.

These are a few of the new developments announced recently. They are examples of the kinds of improvements many quartz crystal oscillator manufacturers are making in their products. Designers are urged to contact all manufacturers for information before specifying a product.

Designers specifying crystal oscillators should be familiar with a recent study by V. Rosati, S. Schodowski and R. Filler of the Electronics Technology & Devices Laboratory, Fort Monmouth, New Jersey. Oscillator vendors were asked to identify two temperature-compensated crystal oscillators for testing, one to meet certain specifications and the other to be their best available TCXO. A total of 105 oscillators were tested for frequency-voltage, frequency-load, frequency-temperature and aging. The researchers asked for moderate (5.0 ppm) or high (0.5 ppm) stability oscillators.

Oscillators in the 1-5 ppm stability range were found to conform to manufacturers' specifications, but long-term aging characteristics were widely variable. A significant percentage of those between 0.5 ppm and 1 ppm were found to be out of specification. No correlation between price and performance was found. The researchers concluded that system designers should carefully specify and test TCXOs before designing them into systems and supporting data should be requested for each oscillator being considered in the 0.5 ppm class.

Reporting the study results, the laboratory's Frequency Control branch chief, J.R. Vig, said "historically the accuracy of quartz oscillators has improved by an order of magnitude approximately every seven to 10 years." He said he expects the rate of improvement to continue.[1]

Vig said he expects revolutionary improvements in TCXOs from the use of microcomputers and "dual-mode" techniques. He said a dual-mode device measures crystal temperature without using an external sensor. The temperature-frequency profile of a crystal is stored in a dedicated microcomputer chip that controls compensation as a function of crystal temperature. Vig predicted frequency-temperature stability of 5×10^{-9} by 1990, with designs that are one-third smaller and use one-third as much power as current TCXOs.

SAW Devices

At frequencies above 100 MHz Surface Accoustic Wave (SAW) devices have a cost and a size advantage over quartz crystal oscillators, which need frequency multipliers, and a frequency stability advantage over LC circuit oscillators.

A typical SAW resonator consists of thin metal strips laid or deposited on a thin substrate, usually of quartz or lithium niobate. High frequency AC voltage across the strips creates surface waves along the face of the crystal. Because of the piezoelectric effect of the crystal the surface waves are detected by other strips on its surface. The principal feature of the SAW resonator is that the surface waves travel many times slower than the electromagnetic waves causing them. This frequency reduction feature allows SAW resonators to handle frequencies up to 1 GHz. Because they can work directly in higher frequencies, SAW oscillators avoid non-harmonic spurious outputs that are a problem with crystals and frequency multipliers.

SAW devices can be designed with taps at any point for multiple use of the oscillating signal. They are often used in hybrid modules and components because they are so small.

The major problems with SAW devices are frequency drift as the crystal substrate ages and contaminants introduced in the manufacturing process. Aging can be compensated for as quartz crystal resonator aging is compensated for. The amount of contamination depends on the care taken when the SAW device is made. SAW devices must be hermetically encased. Water vapor condensing on the surface of the substrate could slow the surface wave and might mix with contaminants to form acids that could etch the surface, changing its wave propagation characteristics.

Trends in SAW oscillator design are toward higher frequencies and better temperature stability. The higher frequencies (mass production above 400 MHz) are aimed at the cellular radio and microwave link markets. Better temperature stability will probably be achieved by using different materials.

Andersen Laboratories uses a SAW delay line as the frequency controlling element in their standard hybrid oscillator. They say their oscillators are capable of operating at fundamental frequencies in the VHF and UHF ranges and beyond, reducing the need for frequency multipliers and post multiplier filtering. The oscillator signal is coupled out of the loop by a power divider and further isolated from the load by the output amplifier, which minimizes load pulling.

Avantek, Inc., is working on development of integrated subassemblies at the substrate level, combining thin-film oscillators, switches and monolithic amplifiers into an oscillating unit that can be tuned over several octaves. They expect to be in production by the end of the year.

Only a few of the many reputable oscillator manufacturers have been mentioned in this report. It was not our intention to present a complete report on all available manufacturers or devices. Such a report would be imcomplete the day it was published because new designs or improved devices are being announced almost weekly. Only by reading new product announcements and literature from manufacturers can the designer keep well informed.

Designers may obtain specific information about currently manufactured oscillators by circling the following Info/Card numbers.

Crystal Oscillators

Austron — Info/Card #130
Aydin Vector — Info/Card #129
Bliley Electric — Info/Card #128
Cinox — Info/Card #127
CTS/Knights Div. — Info/Card #126
Dale Electronics — Info/Card #125
Frequency Control Products — Info/Card #124
Frequency Sources — Info/Card #123
Greenray Industries — Info/Card #122
Hewlett-Packard — Info/Card #121
K&L Quartztek — Info/Card #120
McCoy Electronics — Info/Card #119
Merrimac Industries — Info/Card #118
Microsonics — Info/Card #117
Motorola, Inc. — Info/Card #116
Piezo Crystal — Info/Card #115
Piezo Technology — Info/Card #114
Seiko — Info/Card #113
Signetics — Info/Card #112
Sokol — Info/Card #111
Vectron Laboratories — Info/Card #110
Wenzel — Info/Card #109

SAW and Other Types

Alpha Industries — Info/Card #108
Andersen Laboratories — Info/Card #107
Avantek — Info/Card #106
Crystal Technology — Info/Card #105
Phonon — Info/Card #104
RF Monolithics — Info/Card #103
Sawtek — Info/Card #102
Teledyne — Info/Card #101

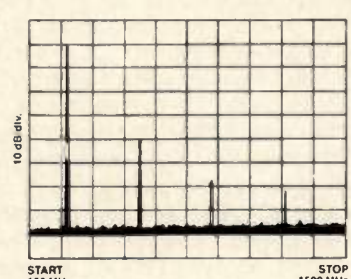
Footnotes

1. Phillip J. Klass, "Advances Forecast in Crystal Oscillators," Aviation Week and Space Technology, Nov. 28, 1983, pp 148-149.

SAW Oscillators

Sawtek's Surface Acoustic Wave oscillators for military and commercial applications simplify design and improve noise performance. High-Q SAW resonators offer quartz stability at fundamental frequencies from 100 MHz to 1250 MHz. Hybrid oscillators in hermetic packages are available for reduced size and increased reliablity. FM or pulse code modulation capability is optional.

Sawtek maintains a large selection of frequencies from an inventory of pre-tooled resonator crystals and new designs can be tooled rapidly. Our engineers also offer assistance in oscillator design for specific applications and are prepared to help evaluate the suitability of SAW oscillators for your requirements.



In addition to oscillators and resonator products, Sawtek produces other high performance SAW components including bandpass filters, delay lines, and pulse compressors for cable television, satellite communications, modems, radar, EW, and many other signal processing applications. And, if what you need is not among our hundreds of standard products, we can provide technical assistance and rapid response to new design and production requirements. Quality and performance have made Sawtek the industry leader in SAW technology; you can rely on us for the total engineering support you need.

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INFO/CARD 17

Controlling Radiated EMI From Digital Systems

By W. Scott Bennett
Hewlett-Packard Company
Ft. Collins, Colo.

Radiated electromagnetic interference (EMI) from digital systems results from the poor design — or lack of design — of current-path geometries in digital systems. In digital circuits, if the path followed by each return current is not effectively the same as the path followed by its associated forward current and the current has RF Fourier components, unacceptable levels of radiated EMI will almost certainly result. This can be predicted from the expressions for the electric and magnetic fields of a small loop antenna(1), both of which are directly proportional to the area bounded by the loop. Current loops in a digital system are not all necessarily "small;" however, the fields of any large current loop can be equivalently represented as the superposition of the fields from a number of smaller current loops having the same perimeter as the large loop. Thus, to effectively control radiated EMI from digital systems one must effectively reduce the areas bounded by the current loops in such systems.

A coaxial cable, for example, can make the effective area enclosed by its current loop equal to zero. At every point external to the cable there are two components which make up the electric field and two components which make up the magnetic field. In each case, one component is due to the (forward) current in the inner conductor and one component is due to the (return) current in the outer conductor. However, because the conductors are coaxial the field components result from currents which are coaxial as well. The forward and return currents appear to be following the same path but they are oppositely directed. If, at each point on the path the currents are *equal* and oppositely directed, at each point outside the cable the net fields due to the currents are zero.

A simple method for reducing current loop areas on a printed circuit (PC) board is to make one full layer of the board a ground plane. Return currents will then tend to follow the same paths in the ground plane as their associated forward currents follow in adjacent layers. The distance between each forward current path and its return current path will then be the distance between the layer of the forward current trace and the layer of the ground plane. Thus, all current loop areas on the board will tend toward zero and radiated EMI due to those loop areas will tend toward zero as well. This same technique can be applied to flat multi-conductor cables which connect one PC board to another.

To obtain a quantiative understanding of the importance of minimizing current loop areas, consider the following equation:

$$A = 58.9 \frac{Ed}{If^2}$$

in which A represents current loop area in square inches, E represents electric field strength in microvolts/meter, d represents distance from the loop in meters, f represents the frequency of a current component in Megahertz and I represents the amplitude of a current component in milliamperes. This equation was derived from the expression for the maximum electric field of a small current loop, assuming the loop is positioned over a large conductive ground plane. It provides a good worst-case approximation of the electric field that would be observed from such a loop when measured according to FCC specifications(2).

Now, suppose one wishes to know the loop areas which are associated with the FCC Class A and Class B radiated EMI limits for digital devices at 200 MHz. At that frequency the Class A limit is an electric field strength of 50 μV/m measured over a ground plane at a distance of 30 meters and the Class B limit is an electric field strength of 150 μV/m measured over a ground plane at a distance of 3 meters(2), (3). From the above equation, the loop area associated with the Class A limit for a 1 mA current component at 200 MHz is 2.2 in^2 and the loop area associated with the Class B limit for the same current and frequency is 0.66 in^2. It should be clear from the small sizes of these areas that current loop areas *must* be minimized by digital designers if the FCC regulations are to be met.

Finally, since the current loop areas associated with the FCC radiated EMI limits for digital systems are so small, it should be clear that long cables between major components of digital systems must be designed *very* carefully. For example, any common mode currents measured on a multiconductor cable of more than a few inches in length will more than likely be a problem because the associated return current paths must lie elsewhere. Although current components in I/O cables, for example, are generally lower in frequency than current components on PC boards, current loop areas established by I/O cables are generally much larger than those of PC boards because the cables are generally long and the PC boards are relatively small. Thus, from long intrasystem cables one generally has radiated EMI problems at frequencies lower than those of PC board radiation.

In summary, the importance of RF current-path geometries cannot be overlooked if one is to design marketable digital systems.

1. J.D. Kraus, *Antennas.* New York: McGraw-Hill, 1950, p.157.
2. U.S. Federal Communications Commission, "Methods of measurement of radio noise emissions from computing devices," in *Rules and Regulations,* pt. 5, subpt. J. Appendix A, pp. 163-171, July, 1981.
3. R.K. Keenan, *Digital Design for Interference Specifications.* Vienna, VA.: The Keenan Corp., 1983, p. A1-5.

The SAW Resonator Filter: How It Works. Part II

By Jeff Schoenwald
Contributing Editor

Last month, SAW resonator theory and a program for the TI-59 calculator and printer were described. This month, we use the program to learn about SAW resonators. Of special value is the definition and order of parameter entry.

Unless otherwise indicated, certain initial parameters that must be entered will always be the same:

$l_1 = l_2 = 0.25$
$n_2 = l$
$f_o = 100$

As a first example, consider the characteristics of a single reflector array as a function of changing strip/gap impedance ratio Z at a fixed number of segments N = 200. Figures 1a and 1b show the reflection loss and transmission power loss respectively for Z = 1.005, 1.01 and 1.015. At frequencies above or below f_c the reflectors are transparent to traveling waves, so the transmission loss is small and the reflected wave loss is quite large. Closer to center frequency, the reflection efficiency of the gratings improves and the transmission loss increases relative to the impedance mismatch Z or number of segments in the array.

In the next example, shown in Figure 2, two reflectors are cascaded. Z = 1.01, N = 100 in each reflector, and the cavity is 5 wavelengths long (i.e., $l_g = 4.75$). Away from the center frequency, the results are similar to that of the previous examples. Near center frequency, the Fabry Perot cavity becomes resonant and the structure is "transparent," and at f_c the reflected energy is negligible.

While we have assumed no losses due to absorption in the transmission line, we will still never obtain an infinite Q factor in a lossless device because a reflective array can never be 100% efficient unless Z or N go to infinity. What we do observe is a resonant peak at the center frequency with the 3 dB fractional bandwidth that determines the unloaded Q ($= f_c \Delta f_{3\ dB}$) of the resonator. Figure 3 shows the power transmission characteristics near $f_o = f_c$ for a single cavity resonator (M = 2), with 100 sections/reflector, Z = 1.01, with the cavity size as the variable. It is apparent that the 3 dB bandwidth of the resonance decreases and Q increases as the cavity length increases. Figure 4 shows how the Q varies with

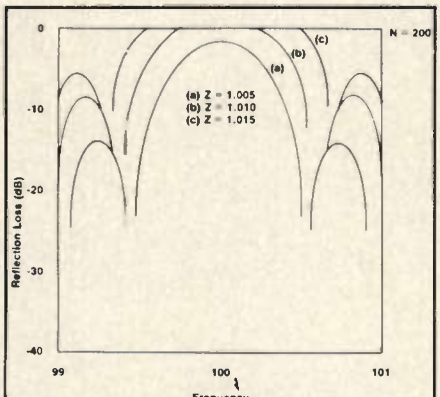


FIGURE 1a. Dependence of SAW grating reflection loss on relative impedance Z.

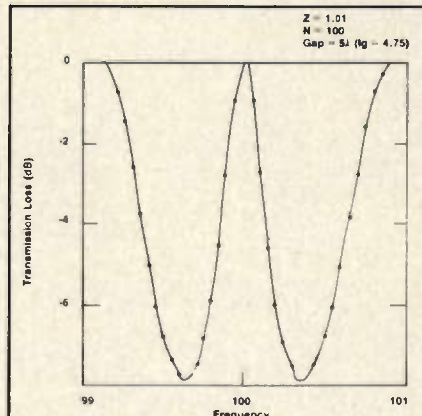


FIGURE 2. Single pole (M = 2) resonator

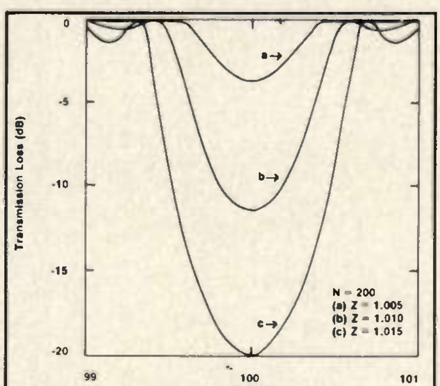


FIGURE 1b. Dependence of SAW grating transmission loss on relative impedence Z.

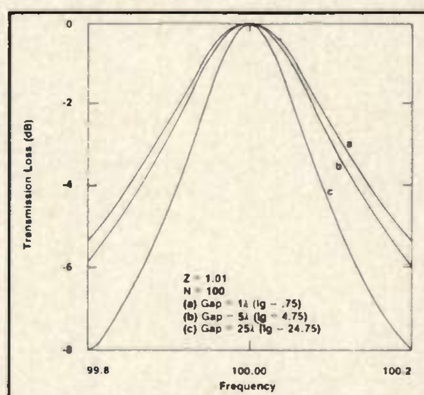


FIGURE 3. Single Pole Resonator (M = 2) dependence on resonant cavity size.

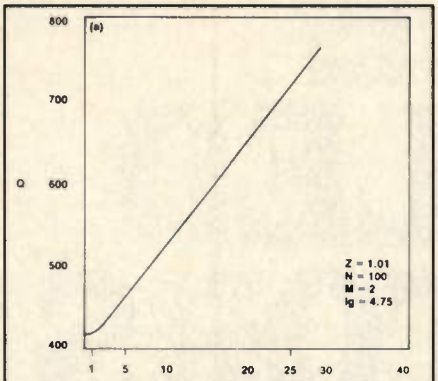


FIGURE 4. Qu dependence on cavity length at fixed Z.

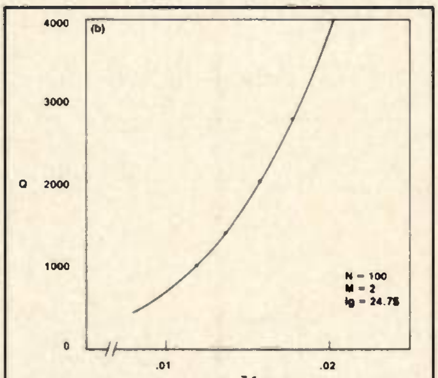


FIGURE 5. Q dependence on impedance mismatch Z-1 at fixed lg.

cavity length while N and Z are held constant. Similar sets of curves can be obtained to determine the dependence of Q or N or Z, which I encourage you to do. Figure 5 shows how Q varies with Z at fixed cavity size.

Cavity Dimension Is Critical

The resonant Fabry-Perot cavity formed by the two reflective arrays will act as an interference filter that effeciently transmits energy at wavelengths which satisfy the cavity relationship:

$$L = \frac{n\lambda_c}{2} \qquad (1)$$

where $L = (l_1 + l_g) \bullet \lambda c$. λc is the normalized wavelength at the peak of reflectivity of the grating. If the wavelength is different from λc, narrow band transmission through the structure will not occur at the center of the reflection stop-band of the grating. Another way of saying this is, if we let $L = l_1 + l_g$ be equal to some non-integer multiple of $\lambda c/2$, the transmission spectrum appears non-symmetric. Figure 6 illustrates the effect of advancing the cavity dimension successively by $\lambda c/8$. If resonance is not precisely centered at the stop band center of the reflectors the Q, selectivity and efficiency of the resonator suffer.

Another interesting case is worth mentioning before we move on to more complicated structures. If we obey the Fabry-Perot condition of Eq. (1), but make n very large, we encounter a regime in which several longitudinal modes of the cavity can. exist, all at slightly different wavelengths (frequencies). This is precisely the case for lasers. A simple laser with two mirrors will have many possible longitudinal cavity modes because the cavity is orders of magnitude larger than wavelength. Special techniques are used to lock the laser into a single mode for added spectral purity. Figure 7 shows some examples of multiple resonances due to oversized cavities.

Designing Multi-Pole Filters

The program is capable of synthesizing device designs that exhibit the essential features of multi-pole SAW resonator filtering behavior. There is no limit to the number of poles one may wish to model, nor does this affect the speed at which the frequency response is synthesized. A design consisting of M grating reflectors will result in an M-1 cavity structure with an M-1 pole response. This assumes, of course that the cavities are kept small enough to avoid more than a single longitudinal mode in each one. The resonances in each cavity will begin to couple and we shall begin to see how this coupling through the grating reflectors that separate each cavity give rise to the very narrow bandpass filtering which has made the SAW resonator popular in the RF and signal processing community.

With this program, all cavities will be the same size and all reflectors will have the same number of strip/gap segments for any one example run. How much can you expect from the TI-59? *C'est la vie.* We will, however, squeeze a bit more out of it.

Figures 8a, b, c and d show the transmission characteristics of 1-, 2-, 3-, and 4-pole resonators. All reflectors have 200 sections, with Z = 1.01, and all cavities are 5λc long (i.e., $l_g = 4.75$). The valleys between resonant peaks are very characteristic of the SAW resonator multi-pole filter. Typically, interdigital transducers (idts) are placed within the two outermost cavities of such a structure. This has the property of enhancing the resonance behavior over the broadband response of the idts, since by themselves they have an insertion loss that is often 20 or 30 dB,

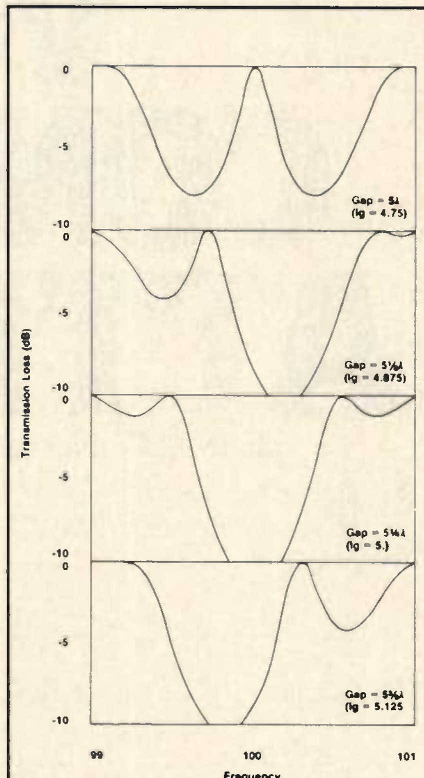


FIGURE 6

whereas the resonances are much stronger because the device impedance drops markedly at those frequencies. Before the idt is properly matched electrically, the resonator response is essentially that of the "unloaded" device and is quite well reproduced by the TI-59 program model. Impedance matching the electrical equivalent circuit of the idt to the cavity impedance will flatten the transmission response over the multipole passband.

The Phase of the Reflected and Transmitted Waves

Once the transmission matrix for the entire structure has been calculated, we may easily determine the phase of the reflected and transmitted wave from the transmission factor and reflection factor, T_F and R_F, using equations 2a and 2b.

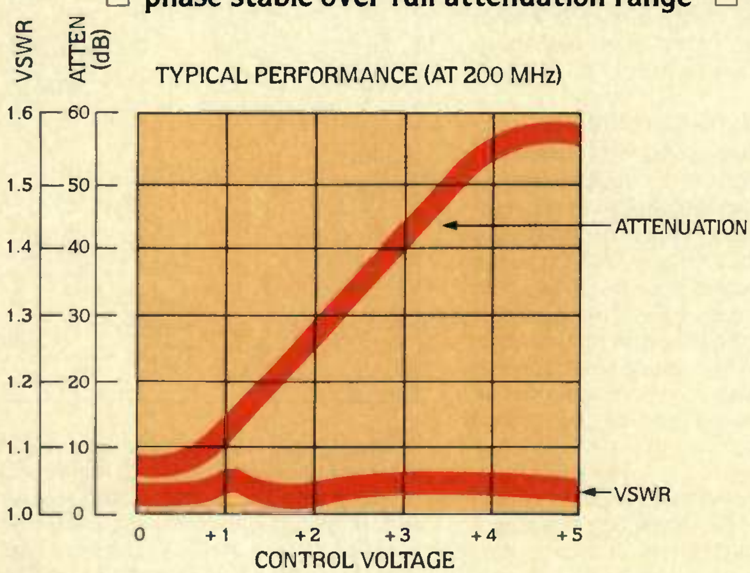
$$\Phi(\text{refl}) = \arctan \frac{\text{Imag}(Rf)}{\text{Real}(Rf)} \qquad (2a)$$

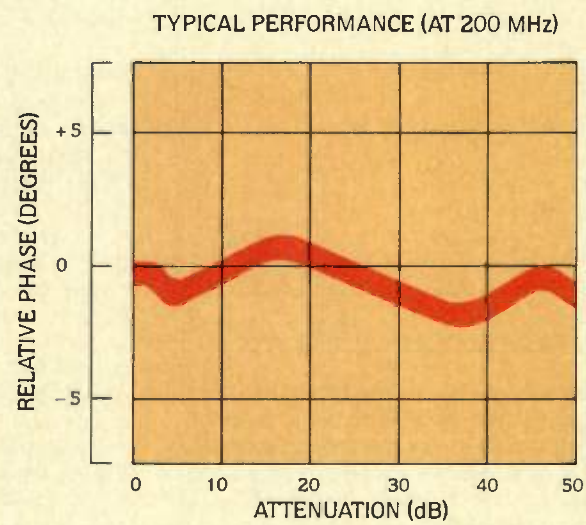
$$\Phi(\text{trans}) = \arctan \frac{\text{Imag}(Tf)}{\text{Real}(Tf)} \qquad (2b)$$

PHASE COMPENSATED

VOLTAGE CONTROLLED ATTENUATORS

☐ phase stable over full attenuation range ☐ internal linearizer ☐ internal driver





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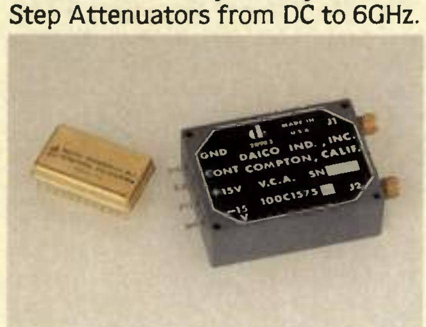

TYPICAL SPECIFICATIONS

frequency	150-250MHz
phase stability	10° max over specified attenuation
attenuation range	40 dB min
linearity	±2 dB
insertion loss	4 dB max
VSWR	1.25 max
control	0 to +5 volts
RF power	+15 dBm max
DC power	+15 volts @ 40 mA −15 volts @ 40 mA
impedance	50 ohms
size	1.5 × 2.0 × .6 in
connectors	SMA
part number	100C1575

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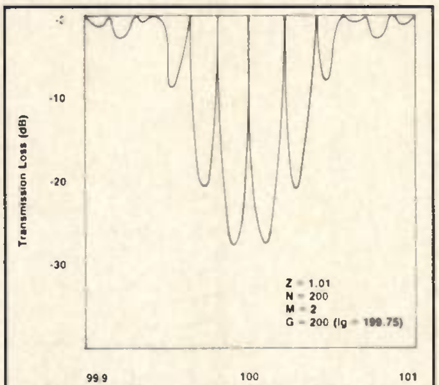

FIGURE 7 frequency

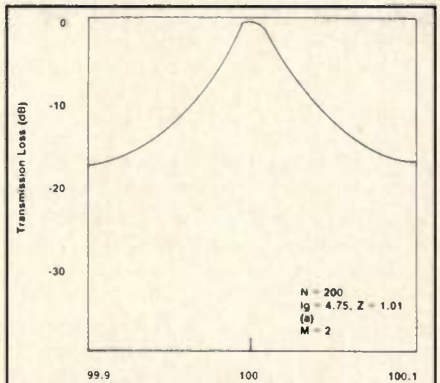

FIGURE 8(a). Single pole SAW resonance (M = 2)

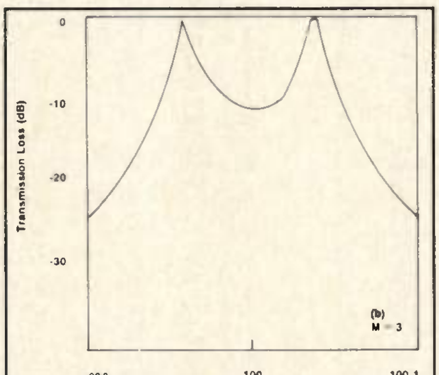

FIGURE 8(b). Two poles (M = 3)

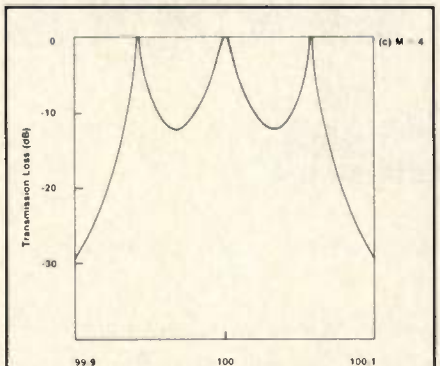

FIGURE 8(c). Three poles (M = 4).

In order to implement this calculation, the listing shown in Table I may be substituted for the last portion of the program, starting at step 638. This may be stored in a simple way. Since the calculator is partitioned for 720 steps, the entire program can be stored in three data banks. The modification occurs only in bank 3. The modified version can be stored on the second edge of the second card (normally used for bank 4). Be sure the calculator is in floating point mode (exit the learn mode and press INV 2nd FIX). Press 3 2nd WRITE but feed in the second magnetic card upside down. Later, when loading the program for phase analysis, the calculator will recognize this as a bank 3 read.

Using this method, power magnitude calculations can be done first, then the third bank replaced with phase calculation, the input paratmeters re-entered and the calculation repeated, yielding the phase response. Not all registers containing the input parameters are re-used. Look at Table II to determine at which label you must re-enter data. Practically speaking, however, it is easier and safer to re-enter all the data from scratch.

Figures 9 and 10 are examples of the phase response for a single pole resonator. They have identical structures but different values of Z. Rather than shortchange the phase computational ability of this program, I will spend a few moments to explain some of the things you can learn from it about SAW resonators. For a plane wave incident on a reflecting surface, the general form of the reflected wave, relative to an incident wave of unit amplitude and zero phase at the front of the reflector, is

$$R_f = re^{2ik \cdot D} \quad (3)$$

where r is the coefficient of reflection magnitude and D is the distance from the point at which the observation is made to the plane of the reflecting surface. If we imagine that the observation point is the front edge of the distributed grating reflector and D is the distance to a fictitious point at which we may assume a single reflecting mirror is placed, then we are constructing a simple model that represents the distance to the "effective center of reflection."

The phase of the reflected wave is given by the argument of the exponential. If we take the first derivative of this phase with respect to the frequency and convert from radians to degrees, we obtain

$$\Phi_{rad} = 2\frac{2\pi}{\lambda}D = \frac{4\pi Df}{\nabla}$$

$$\frac{\delta\Phi_{rad}}{\delta f} = 4\pi\frac{D}{\nu} = \frac{4\pi D}{f\lambda} \quad (4)$$

Therefore,

$$D = \frac{\lambda f}{4\pi}\cdot\frac{\delta\Phi}{\delta f} = \frac{\lambda f}{4\pi}\,\frac{\pi}{180}\,\frac{\delta\Phi deg}{\delta f}$$

$$\frac{D}{\lambda} = \frac{f}{720}\,\frac{\Delta\Phi}{\Delta} \quad (5)$$

Thus, by measuring the slope of phase data we obtain for the reflected wave we are able to compute the distance to the "effective center of reflection." At the peak in reflectivity the center of reflection appears to move toward the front of the reflective array, as evidenced by a decrease in the magnitude of the phase slope. Away from resonance, where the array ultimately looks transparent, the center of reflection moves to the geometric center of the array.

For the cases shown in Figures 10a and 10b, the reflection centers are located, at resonance, a distance of 22 and 17 wavelengths, respectively, inside the array, as viewed from the direction of incidence. Now consider two identical reflectors separated by a distance necessary to produce a single pole resonance. The total length of the cavity is then given by

$$L = (l_g + 1/4) \bullet c + 2D \quad (6)$$

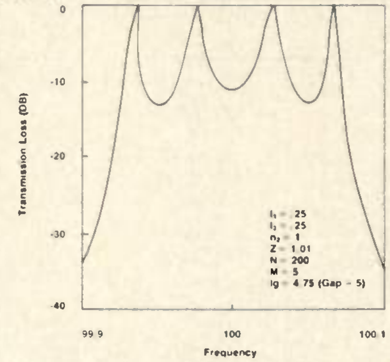

FIGURE 8(d). Four Poles (M = 5)

TABLE I Phase Calculation

638	60	DEG	678	53	(
639	53	(	679	43	RCL
640	02	2	680	06	06
641	65	x	681	85	+
642	53	(	682	43	RCL
643	43	RCL	683	07	07
644	08	08	684	54	)
645	65	x	685	55	÷
646	43	RCL	686	53	(
647	06	06	687	43	RCL
648	75	-	688	05	05
649	43	RCL	689	85	+
650	05	05	690	43	RCL
651	65	x	691	08	08
652	43	RCL	692	54	)
653	07	07	693	94	+/−
654	54	)	694	22	INV
655	55	÷	695	30	TAN
656	53	(	696	95	=
657	43	RCL	697	99	PRT
658	05	05	698	98	ADV
659	33	x^2	699	43	RCL
660	85	+	700	13	13
661	43	RCL	701	32	$X\rightleftarrows T$
662	06	06	702	43	RCL
663	33	x^2	703	26	26
664	75	-	704	67	EQ
665	43	RCL	705	19	D'
666	07	07	706	43	RCL
667	33	x^2	707	29	29
668	75	-	708	42	STO
669	43	RCL	709	11	11
670	08	08	710	61	GTO
671	33	x^2	711	14	D
672	54	)	712	76	LBL
673	54	)	713	19	D'
674	22	INV	714	43	RCL
675	30	TAN	715	26	26
676	95	=	716	91	R/S
677	99	PRT			

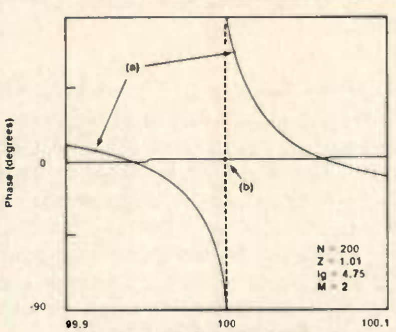

FIGURE 9. (a) Reflection phase and (b) transmission phase of a single pole (M = 2) SAW resonator structure.

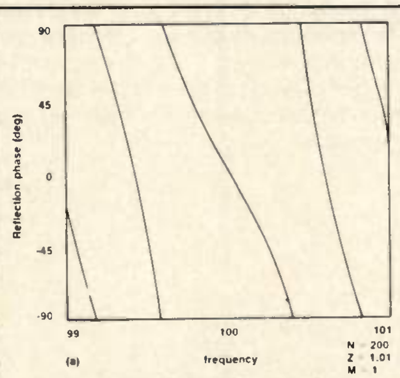

FIGURE 10(a). Single reflective array reflection phase, with N = 200, M = 1, Z = 1.01

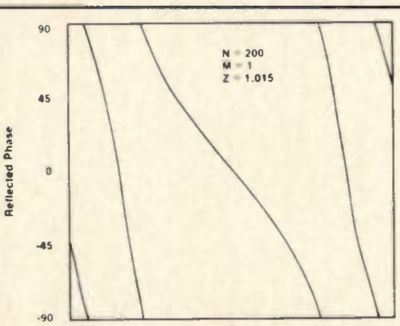

FIGURE 10(b). Single refelctive array reflection phase, with N = 200, M = 1, Z = 1.015.

Knowing the value of D is important for determining the frequency spacing between longitudinal modes in the resonant cavity formed by the two arrays. Now we can see that mode spacing is a function of both cavity size and reflective properties of the array, as represented by the distance D. It is possible to derive frequency spacing depedence on effective length of the cavity L, and such derivations abound in books on lasers. The reader is encouraged to do this for the specific case shown in Figure 7. You will indeed find that the mode spacing observed in large SAW resonator cavities, or in SAW resonators with Z so close to unity that D is sizable, is properly described by taking this reflection distance into account properly. More than once, ignorance of this fact has bitten SAW resonator designers when they least expected it.

Conclusion

I have tried to give the reader a tool for modeling and exploring the behavior of the SAW resonator, a very interesting tool in RF technology. Even if you have no plans for designing such devices you may very likely wish to specify one from a number of vendors that supply them. In that case, you will be better off for having become familiar with the properties of these devices and being an educated customer.

TABLE II

Register	Variable
0	N, N-1, M, M-1
1	λ1, λ2
2	M
3	lg
4	Θ, λ1 ↑N, λ3 ↑N
5	l_1, A(N), A(N + 9)
6	l_2, B(N), B(N + 9)
7	n_2, C(N), C(N + 9)
8	f_0, D(N), D(N + 9)
9	fc
10	Z
11	N
12	t, cos Θ = (A + D)/2
13	f_1
14	df
15	α = Pi * f_1/f_c
16	Z + 1
17	Z – 1
18	1 – 2t
19	A(1), A(N)
20	B(1), B(N)
21	C(1), C(N)
22	D(1), D(N)
23	[((A + D)/2) ↑2 –1]
24	λ1
25	λ2
26	f_2 + df
27	sin NΘ or (λ1 ↑N – λ2 ↑N)/ (λ1 – λ2)
28	sin (N – 1)Θ or (λ1 ↑(n – 1) – λ2 ↑(N – 1))/(λ1 – λ2)
29	N

BONUS!: A *BASIC* Version

For readers who own or work with a computer, here is a BASIC version of the TI-59 calculator program to model the transmission and reflection power and phase of a SAW resonator type of structure. This program was written on a Commodore 64, which uses a version of BASIC and an ASCII character set that is slightly different from standard ones. I have taken pains to write the program to be as transportable as possible, indicating, when necessary, what lines are not transportable and require change or deletion. The POKEs only pertain to screen or letter color. There are no PEEKs. The printer commands are usable "as is" only on Commodore printers and serve as an example which must be tailored for other brands.

Because memory is not as tight for most micros, power and phase are computed and provisions are made for screen display and hard copy. Disk or tape storage of data files and high resolution plotting are routines that the devoted hacker should add, but these have been left out because of differences between computers in storage procedures and screen addressing.

Data entry is simple and screen prompted. If the velocity in a strip segment is different from the gap velocity the center frequency will be different from the designated reference frequency. I suggest you start with a ratio of 1 and then try small changes to observe the trend. The same advice holds for the impedance ratio and cavity size.

```
10 REM  **********************************
20 REM  *      CERTAIN LINES ARE         *
30 REM  *       SPECIFIC TO THE          *
40 REM  *         COMMODORE 64.          *
50 REM  *  DELETE OR ALTER AS DESIRED.   *
60 REM  **********************************
70 PI= 3.14159265:REM   * C-64 RECOGNIZES SYMBOL FOR PI.  USE IT IF YOU CAN *
80 POKE 53281,0:POKE 53280,0:REM            *SET BACKGROUND TO  BLACK FOR C-64*
90 PRINT CHR$(158):REM                     *SET SCREEN LETTERS TO YELLOW ON C-64*
100 PRINT CHR$(147)
110 PRINT"SAW RESONATOR TRANSMISSION"
120 PRINT"     LINE SIMULATION"
130 PRINT""
140 PRINT"(C)OPYRIGHT JEFF SCHOENWALD"
150 PRINT"          1984"
160 FOR I = 1 TO 2000
170 NEXT I
180 PRINT CHR$(147)
190 INPUT"GAP SEGMENT LENGTH, L1";L1
200 INPUT"LINE SEGMENT LENGTH, L2";L2
210 INPUT"RELATIVE VELOCITY INDEX OF L2";N2
220 INPUT"REFERENCE FREQUENCY F0";F0
230 INPUT"STRIP/GAP IMPEDANCE RATIO, Z";Z
240 INPUT"# OF SEGMENT PAIRS PER REFLECTOR, N";N
250 REM
260 INPUT"START FREQUENCY, F1";F1
270 INPUT"FREQUENCY INCREMENT, DF";DF
280 INPUT"STOP FREQUENCY, F2";F2
290 INPUT"# OF REFLECTOR ARRAY SECTIONS, M";M
300 INPUT"CAVITY SIZE, LG";LG
310 FC=F0/(2*(L1+N2*L2))
320 PRINT CHR$(147)
330 PRINT"CENTER FREQUENCY =";FC
340 T=N2*L2/(L1+N2*L2)
350 K1=INT((F2-F1)/DF)
360 DIM AN(K1),BN(K1),CN(K1),DN(K1),TP(K1),RP(K1),TF(K1),RF(K1)
370 FOR K = 0 TO K1
380 F = F1 + DF*K
390 LET GAMMA =PI*F/FC
400 A1=((Z+1)/2)*COS(GAMMA)-((Z-1)/2)*COS(GAMMA*(1-2*T))
410 B1=((Z+1)/2)*SIN(GAMMA)-((Z-1)/2)*SIN(GAMMA*(1-2*T))
420 C1=((Z+1)/(2*Z))*SIN(GAMMA)+((Z-1)/(2*Z))*SIN(GAMMA*(1-2*T))
430 D1=((Z+1)/(2*Z))*COS(GAMMA)+((Z-1)/(2*Z))*COS(GAMMA*(1-2*T))
440 N0=N
450 GOSUB  650:REM *COMPUTE EIGENVALUES FOR ONE ARRAY AND COMPUTE T  MATRIX*
460 GOSUB  970:REM                               *COMPUTE CAVITY T MATRIX*
470 GOSUB 1030:REM                                 *MULTIPLY MATRICES*
480 GOSUB 1090:REM                                *REDEFINE NET T MATRIX*
490 N0=M
500 GOSUB  640
510 REM                                   *COMPUTE EIGENVALUES AND MATRIX FOR*
520 REM                                   *M CASCADED REFLECTORS AND CAVITIES*
530 AN(K)=AN
540 BN(K)=BN
550 CN(K)=CN
560 DN(K)=DN
570 PRINT CHR$(147)
580 PRINT"FREQUENCY PT. #";K;" = ";F
590 GOSUB 1180:REM                                  *COMPUTE POWER & PHASE*
600 NEXT K:REM                                      *NEXT FREQUENCY POINT*
610 REM  *GOTO OUTPUT MENU*
620 GOTO 1760
630 REM
640 REM**REAL OR COMPLEX EIGENVALUES?**
650 TRIG = (A1+D1)/2
660 IF (TRIG^2-1<0) THEN     690
670 IF (TRIG^2-1>0) THEN    870
```

```
680 RETURN
690 REM T MATRIX FOR COMPLEX EIGENVALUEREADY.
700 AGL = -ATN(TRIG/SQR(-TRIG*TRIG+1))+(PI/2)
710 IF AGL=0 THEN    780
720 N1=N0-1
730 AN=A1*SIN(N0*AGL)/SIN(AGL)-SIN(N1*AGL)/SIN(AGL)
740 BN=B1*SIN(N0*AGL)/SIN(AGL)
750 CN=C1*SIN(N0*AGL)/SIN(AGL)
760 DN=D1*SIN(N0*AGL)/SIN(AGL)-SIN(N1*AGL)/SIN(AGL)
770 GOTO   790
780 GOSUB 1210
790 RETURN
800 REM *WHEN EIGENVALUES  ARE*
810 REM          *DEGENERATE*
820 AN=A1-1
830 BN=B1
840 CN=C1
850 DN=D1-1
860 RETURN
870 REM*T MATRIX FOR REAL EIGENVALUES*
880 K1=TRIG+SQR(TRIG^2-1)
890 K2=1/K1
900 N1=N0-1
910 AN=A1*(K1^N0-K2^N0)/(K1-K2)-(K1^N1-K2^N1)/(K1-K2)
920 BN=B1*(K1^N0-K2^N0)/(K1-K2)
930 CN=C1*(K1^N0-K2^N0)/(K1-K2)
940 DN=D1*(K1^N0-K2^N0)/(K1-K2)-(K1^N1-K2^N1)/(K1-K2)
950 RETURN
960 REM  ***GENERATE CAVITY MATRIX***
970 REM
980 A=COS(2*GAMMA*LG)
990 B=SIN(2*GAMMA*LG)
1000 C=SIN(2*GAMMA*LG)
1010 D=COS(2*GAMMA*LG)
1020 RETURN
1030 REM ***MATRIX MULTIPLICATION***
1040 AP=AN*A-BN*C
1050 BP=AN*B+BN*D
1060 CP=CN*A+DN*C
1070 DP=DN*D-CN*B
1080 RETURN
1090 REM
1100 A1=AP
1110 B1=BP
1120 C1=CP
1130 D1=DP
1140 RETURN
1150 REM ***COMPUTE REFLECTED AND***
1160 REM ***TRANSMITTED POWER AND***
1170 REM       ***PHASE***
1180 Q= (AN(K)+DN(K))^2+(BN(K)+CN(K))^2
1190 TP=4/Q
1200 TP(K)=10*(LOG(TP))/LOG(10)
1210 RP=1-TP
1220 IF RP = 0 THEN   1280
1230 RP(K)=10*(LOG(RP))/LOG(10)
1240 TF(K)=(180/PI)*ATN(-1*(BN(K)+CN(K))/(AN(K)+DN(K)))
1250 RF=2*(BN(K)*DN(K)+AN(K)*CN(K))/(AN(K)^2-DN(K)^2+BN(K)^2-CN(K)^2)
1260 RF(K)=(180/PI)*ATN(RF)
1270 RETURN
1280 RP(K) = -99.0
1290 GOTO 1240
1300 W=20:REM                          *STOP SCREEN PRINT EVERY 20 LINES*
1310 PRINT CHR$(147)
1320 PRINT"FREQ.  R-PWR  T-PWR(DB)  R-PHA  T-PHA"
1330 PRINT"        (DB)      (DB)      (DEG)    (DEG)"
1340 FOR K=0 TO K%
1350 F= F1+DF*K + .000001
1360 B$=STR$(F):GOSUB 1690:F$=B$
1370 R1$=STR$(RP(K))
1380 B$=R1$:GOSUB 1590:R1$=B$
1390 T1$=STR$(TP(K))
1400 B$=T1$:GOSUB 1590:GOSUB 1670:T1$=B$
1410 R2$= STR$(RF(K))
1420 B$=R2$:GOSUB 1590:R2$=B$
1430 T2$= STR$(TF(K))
1440 B$=T2$:GOSUB 1590:T2$=B$
1450 IF W=K%+2 THEN 2090
1460 PRINT F$;TAB(14-LEN(R1$));R1$;TAB(22-LEN(T1$));T1$;TAB(30-LEN(R2$));
1470 PRINT R2$;TAB(39-LEN(T2$));T2$
1480 IF K=0 THEN   1510
1490 IF K=K%THEN   1520
1500 IF K/W =INT(K/W)THEN   1920
1510 NEXT K
1520 IF W=K%+2 THEN 2050
1530 PRINT"END OF LISTING.  PRESS C FOR MENU"
1540 GET A$
1550 IF A$=""THEN   1540
1560 IF A$="C"THEN   1760
1570 IF A$<>"C"THEN   1540
1580 RETURN
1590 REM                          *MANIPULATE STRINGS TO LINE UP COLUMNS*
1600 FOR U = 1  TO  LEN(B$)
1610 IF MID$(B$,U,1)="."THEN   1640
1620 NEXT U
1630 RETURN
1640 B$=LEFT$(B$,U+2)
1650 IF LEN(RIGHT$(B$,U+1))<2 THEN B$=B$+"0"
1660 RETURN
1670 IF ABS(TF(K))<.01 THEN B$="-0.00"
1680 RETURN
1690 FOR U = 1  TO  LEN(B$)
```

```
1700 IF MID$(B$,U,1)="." THEN 1730
1710 NEXT U
1720 RETURN
1730 B$=LEFT$(B$,U+3)
1740 IF LEN(RIGHT$(B$,U+1))<3 THEN B$=B$+"00"
1750 RETURN
1760 PRINT CHR$(147)
1770 PRINT"     CHOOSE AN OPTION,  PLEASE"
1780 PRINT""
1790 PRINT"1 - REPEAT CALCULATION",CHR$(13)
1800 PRINT"2 - PRINT DATA TO PRINTER",CHR$(13)
1810 PRINT"3 - PRINT DATA TO SCREEN",CHR$(13)
1820 PRINT"4 - QUIT PROGRAM",CHR$(13)
1830 GET A$
1840 IF A$="1"THEN  1890
1850 IF A$="2"THEN  2000
1860 IF A$="3"THEN  1300
1870 IF A$="4"THEN END
1880 IF A$="" THEN  1830
1890 CLR
1900 GOTO 190
1910 PRINT CHR$(147)
1920 PRINT"PRESS (C) TO CONTINUE"
1930 GET A$
1940 IF A$="C" THEN   1960
1950 IF A$<>"C"THEN   1930
1960 PRINT"FREQ.       R-PWR     T-PWR   R-PHA   T-PHA"
1970 PRINT"            (DB)      (DB)    (DEG)   (DEG)"
1980 GOTO 1510
1990 PRINT CHR$(147)
2000 W=N1+2
2010 OPEN 4,4:CMD4   REM   * C-64 COMMAND TO OPEN CHANNEL TO COMMODORE PRINTER *
2020 REM                   * TO SUIT YOUR PRINTER PLEASE ALTER THE FOLLOWING   *
2030 REM                   * LINES 2010, 2050, 2100 AND 2110                   *
2040 GOTO 1330
2050 PRINT#4: CLOSE4
2060 PRINT CHR$(147)
2070 PRINT "END OF HARD COPY LISTING"
2080 GOTO 1770
2090 REM
2100 PRINT#4,F$,CHR$(16)CHR$(49)CHR$(48)R1$,
     CHR$(16)CHR$(49)CHR$(56)T1$,
2110 PRINT#4,CHR$(16)CHR$(50)CHR$(54)R2$,CHR$
     (16)CHR$(51)CHR$(52)T2$
2120 GOTO 1430
```


Erratum:

There was an error in equation (13) as printed in Part I. It should have read:

$$[Tg] = \begin{vmatrix} \cos\gamma & jZ1\sin\gamma \\ \dfrac{j\sin\gamma}{Z1} & \cos\gamma \end{vmatrix}$$

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University of Mississippi, Oxford Campus
Oxford, MS

Information: Dr. Darko Kajfez, Department of Electrical Engineering, University, MS 38677; Tel: (601) 232-7231.

March 19-22
Componentes Electronicos '85

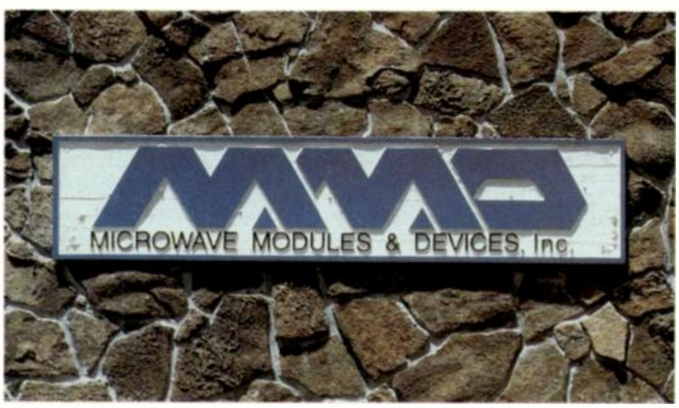
Guadalajara, Mexico

Information: Ms. Raquel Polo, Deputy Project Manager, United States Trade Center, P.O. Box 3087,
Laredo, TX 78044; Tel: (905) 591-0155.

The Company

Microwave Modules & Devices is uniquely dedicated to integrating device and circuit technology to break through current technological barriers to achieve a new level of pricing and performance in the RF/Microwave field.

The professionals who head the company have been engaged in the design, development and production of high power, solid-state microwave modules and devices for over 15 years. Today, MMD's expanding line of products includes amplifiers employing both hybrid micro-circuitry and modular printed circuit technology. The company's high performance amplifiers are targeted for use in military and industrial applications worldwide. In addition, MMD is deeply involved in the design, development and manufacture of a wide variety of RF and microwave functions for high technology applications both for military and industrial markets. MMD's markets include communications, countermeasures, radar, space, broadcast, avionics, industrial, scientific and medical. Reduction in size is one of MMD's on-going goals while constantly striving to improve both performance and product reliability. This brochure highlights MMD's current capabilities and its technological direction in the future.



Custom Modules

MMD specializes in developing and manufacturing modules to your specifications. What do you need to maintain your competitive edge?

- Higher power?
- Wider bandwidth?
- Better efficiency?
- Smaller size?
- Improved reliability?

MMD can solve these design problems to meet your performance requirements.

The Plant

Microwave Modules & Devices (MMD) is located in Mountain View, California, in the heart of the high technology area known internationally as "Silicon Valley." The modern 10,000 square foot plant is fully equipped with the most advanced test, process and assembly equipment. It has clean room and assembly capabilities for delivering softboard and hybrid modules to the demanding requirements of military programs and sophisticated industrial applications. MMD's quality assurance procedures are designed to incorporate the latest military standards.

Fast Response Time

Fast response to customer amplifier needs is assured by a complete in-house design, prototyping and manufacturing capability. This is due totally or in part to:

- Complete engineering model shop.
- Computer-aided design.
- Automated die attach.
- Automated test capabilities.
- Engineering staff with more than 125 years combined experience.
- Computerized manufacturing/accounting systems.

Soft Board Amplifiers

State-of-the-art performance, high reliability, small size and fast development time are among the challenges that MMD is accepting every day. If these sound like your particular problems, MMD can provide the solutions you have been looking for. For example, complete solid-state amplifiers with power outputs to more than 1 KW can be designed and manufactured to your specifications. If you have a new design requirement for TV, radar, communications, avionics, EW, or need a solid state replacement for a vacuum tube, MMD is in business to solve your problems **fast!!** If improvements in efficiency or linearity are important in your application, MMD can help. In fact, MMD is providing an order of performance previously unachievable in RF power amplifier modules. MMD can **demonstrate** how to reach such goals as wide bandwidth, high efficiency, power outputs to

more than 1 KW Class C or Linear, Pulse or CW. All of this can be achieved by several unique MMD capabilities including:

- Custom die designs for optimized performance.
- Custom BeO ceramic chip carriers for best die/circuit interfaces.
- In-house custom ceramic/metal device packaging for lower thermal resistance; lower parisitics; hermetic seal.
- Device assembly.

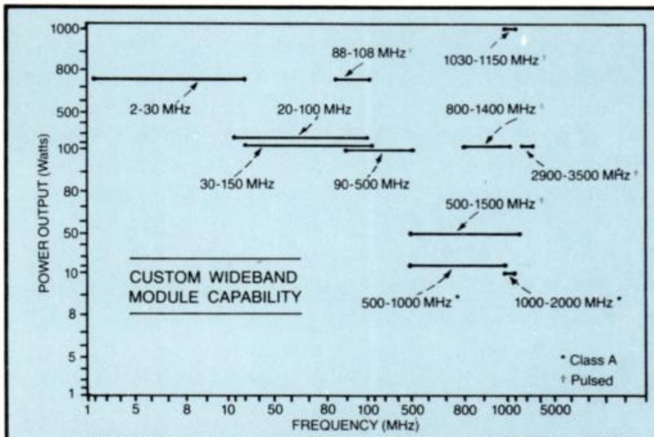
For example, these unique capabilities come together to create MMD's **"POWER QUAD"** concept which combines with new wideband matching components to offer power/bandwidth capability far superior to that of conventional designs. So, if **performance** is an important consideration, call and find out how Microwave Modules & Devices can provide you with a **cost effective** alternative to in-house design, development and manufacturing of your critical power amplifier modules.

RF Functions

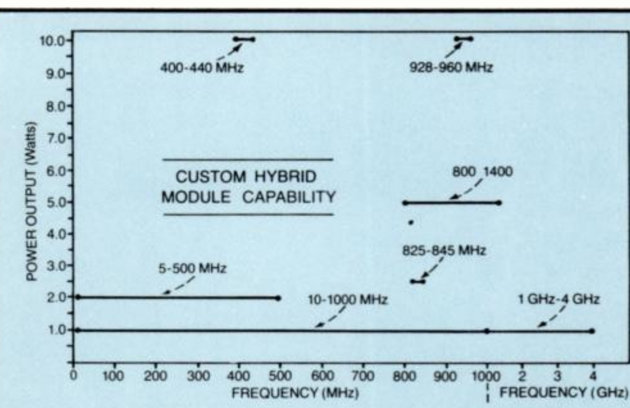
MMD can provide all the RF functions you need for your particular system requirements. Active mixers, AGC amplifiers, high speed modulators, wideband amplifiers and IF amplifiers are just some of the products under development. MMD functions use transistors with Ft up to 8 GHz and power capability up to 1 watt, making them ideal for low distortion, wide dynamic range EW applications. A variety of packages are available for your custom requirements such as welded hermetic enclosures or small microstrip styles.

Hybrid Power Amplifiers

MMD hybrid power amplifier modules using chip transistors and either thick or thin film construction offer improved performance and ease of use compared to discrete designs. Hybrid power amplifier modules can provide guaranteed performance with no additional tuning or impedance matching headaches and the "Drop-in" configuration simplifies the user's mechanical design. In addition, smaller size offers a real advantage in systems where space is at a premium. MMD specializes in custom designs requiring BeO substrates or chip carriers to achieve performance not possible with discrete designs.



FREQUENCY	POWER	FREQUENCY	POWER	FREQUENCY	POWER
2-30 MHz	750 watts	90-500 MHz	125 watts	1000-2000 MHz	10 watts*
20-100 MHz	250 watts	500-1000 MHz	20 watts*	1030-1150 MHz	1000 watts†
30-150 MHz	150 watts	500-1500 MHz	50 watts†	2900-3500 MHz	100 watts†
88-108 MHz	750 watts†	800-1400 MHz	150 watts†		



FREQUENCY	POWER	FREQUENCY	POWER
5-500 MHz	2 watts Class A	800-1400 MHz	5 watts
10-1000 MHz	1 watt Class A	928-960 MHz	10 watts Class AB
400-440 MHz	10 watts Class AB	1 GHz-4 GHz	1 watt Class A
825-845 MHz	25 watts		

PRODUCT BULLETIN No. 53. 1-5-85

MICROWAVE MODULES & DEVICES, Inc.

AC-0003-500W

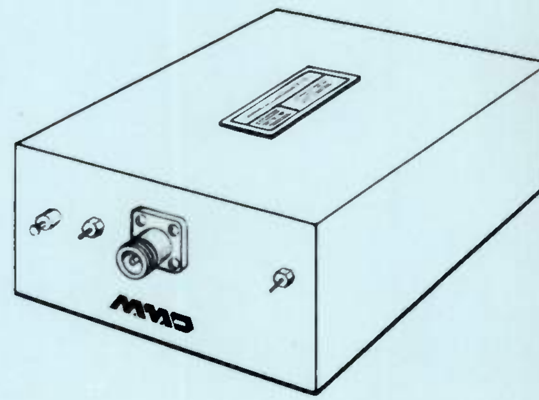
500 WATT SOLID STATE AMPLIFIER MODULE

The AC-0003-500W is a modular wideband HF Class AB Linear Power Amplifier that provides 500 watts of output power at an intermodulation distortion level of better than −35dB (IMD_3). The module incorporates wideband 90° quadrature combiners resulting in predictable 50 ohm system interfaces and superior performance into imperfect loads. The unit also contains internal voltage regulation and reverse voltage protection.

A single AC-0003-500W can be used as the final stage in a 500 watt system or modules can be combined to construct multi-kilowatt transmitters.

The MMD AC-0003-500W module is a reliably rugged building block designed for a wide variety of uses in Military and Industrial HF applications.

2 to 30MHz
HF MODULE



FEATURES:

- HIGH POWER HF BUILDING BLOCK MODULE
- 500 WATT LINEAR OUTPUT
- 50 OHMS IN AND OUT
- WIDEBAND 2 to 30MHz
- 14dB GAIN
- FLAT GAIN VS FREQUENCY
- INTERNAL VOLTAGE REGULATION AND REVERSE VOLTAGE PROTECTION

MODULE SPECIFICATIONS

PARAMETER	LIMITS		UNITS	CONDITIONS
	MIN.	MAX.		
GAIN	13.0		dB	P_{OUT} = 500 WATTS PEP FREQ. = 2 TO 30 MHz V_1 = 5.0 VOLTS V_2 = 45 VOLTS
GAIN FLATNESS		±1.5	dB	
EFFICIENCY	30		%	
INPUT VSWR		1.5:1		
OPERATING LOAD VSWR		2.0:1		
NON-DESTRUCTIVE LOAD VSWR		5.0:1		
IMD_3		−35	dB	
P_{SAT}	700		W	FREQ. = 2 TO 30 MHz V_1 = 5.0 VOLTS V_2 = 45 VOLTS
OUTPUT SOURCE VSWR		1.5:1		
WEIGHT		7	LBS.	

All specifications are at a 30° C baseplate unless otherwise stated.

PRIME POWER REQUIRED: V_1 = 5.0 ±0.1 VOLTS @ 2.5 AMPS.
V_2 = 45.0 ±1.0 VOLTS @ 40 AMPS.

OPERATING BASEPLATE TEMPERATURE: −25°C TO +85°C.

CONNECTORS: TYPE "N" FEMALE (RF). SOLDER PIN (DC).

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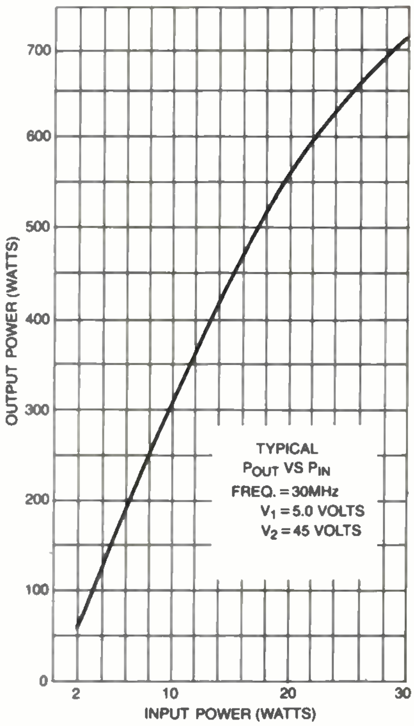
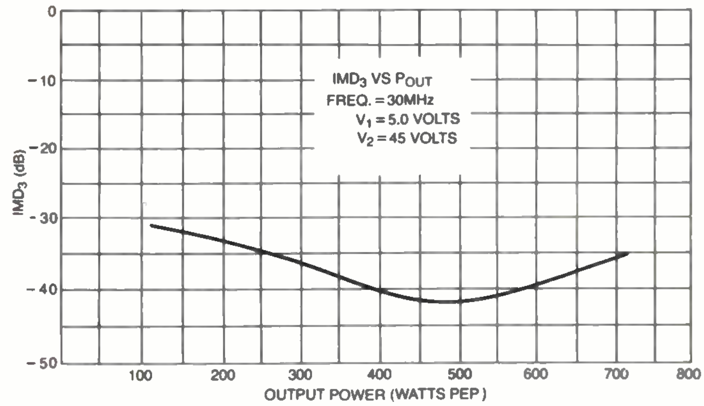
AC-0003-500W

500 WATT SOLID STATE AMPLIFIER MODULE





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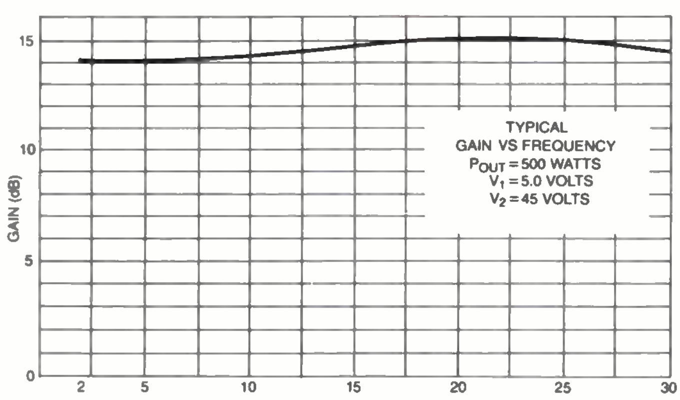
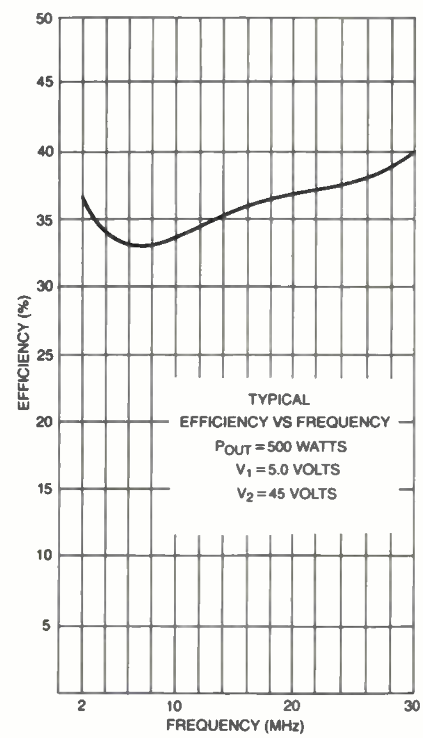
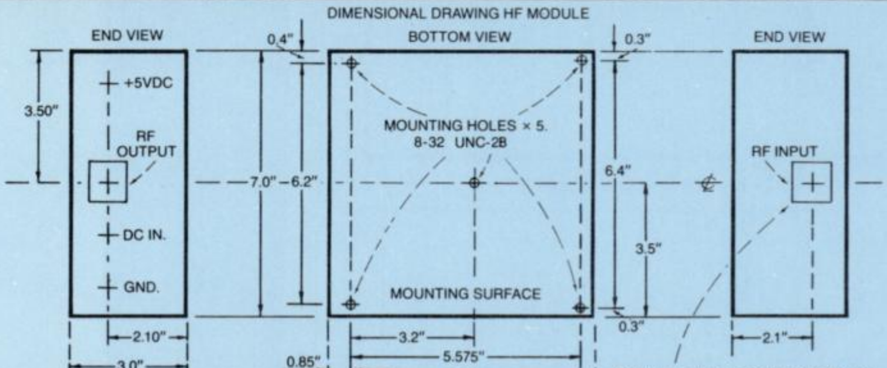




DIMENSIONAL DRAWING HF MODULE



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T-126, 1-8-85, LITHO U.S.A.

PRODUCT BULLETIN No. 52, 12-20-84

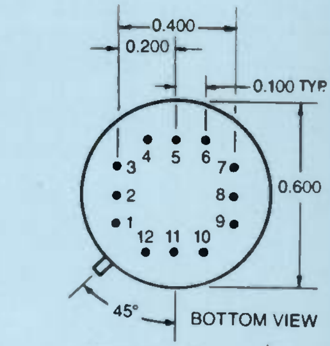
MICROWAVE MODULES & DEVICES, Inc.

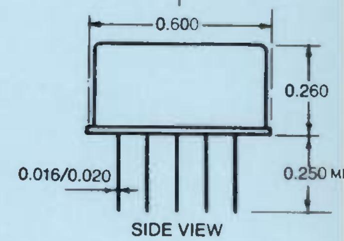
FA-501

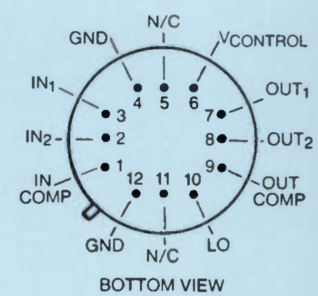
WIDEBAND RF FUNCTION AMPLIFIER

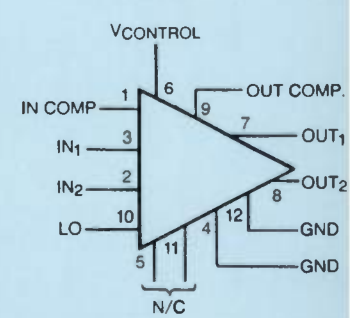
FREQUENCY RANGE: 5-500 MHz.

The FA-501 is an RF function amplifier that offers the circuit designer the versatility of an operational amplifier in both transmitter and receiver applications. With more than 20dB of reverse isolation and unconditional stability, a wide variety of systems problems can be solved. Features include 5.5dB noise figure, 10dB gain, 25dB AGC and a bandwidth that exceeds 500 MHz. In some applications the FA-501 is useful to 750 MHz. The compact unit is housed in a hermetically sealed package.









AMPLIFIER SPECIFICATIONS @ 25°C, V_{cc} = +5V.

Frequency Range:		5-500MHz.
Gain (I_o = 110 ma)		10dB min.
Gain Flatness:	(G = 0 to 10dB)	±1dB
	(G = −15 to 10dB)	±3dB
Power Output @ 1dB Gain Compression (I_o = 110 ma)		+10dBm min.
Noise Figure (I_o = 110 ma)		5.5dB max.
VSWR (50 ohms)	Input:	2.5:1 max.
(G = −10dB to 10dB)	Output:	2.5:1 max.
3rd Order Intercept Point (I_o = 110 ma)		+20dBm typ.
2nd Harmonic Suppression:		−25 dBc @ P_{1dB}
		−40 dBc @ P_{1dB}−10dB.
Reverse Isolation:		20dB min.

AGC RELATED:

Control Range:	5-500 MHz:	25dB min.
	5-100MHz:	30dB min.
Control Input Range:		+1 to +4 V
		2 to 18 ma.
Response Time for 25dBΔ:		1μsec max.
Gain Stability: −30 to +50°C (For Any Setting Within Specified Control Range)		4dB p-p max.

ENVIRONMENTAL:

Operating Case Temp.	−55° to +85°C.
Storage Temp.	−62° to +125°C.

ABSOLUTE MAXIMUM RATINGS:

V_{CC}	+8 volts.
$V_{CONTROL}$ (Without External Resistor)	+5 volts.
Current (I_o):	150 ma.
RF Input Power At Any Port:	+13dBm.

MIXER CHARACTERISTICS:

Frequency Range:	LO:	5-500MHz.
	RF:	5-500MHz.
	IF:	5-500MHz.
Conversion Loss/Gain:	Minimum:	−4dB.
	Typical:	0dB.
	Maximum:	+2dB.

Isolation:	LO-RF:	25dB min.
	LO-IF:	10dB min.
	RF-IF:	0dB min.*
Noise Figure: Typical.	10MHz:	17dB
	200MHz:	17dB
	500MHz:	17dB

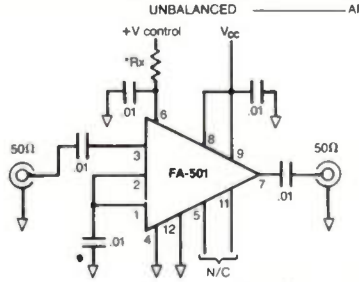
Two Tone IM_3: at −10 dBm RF Port Power	35dBc
VSWR at R Port:	2.0:1 max.
VSWR at L Port:	1.8:1 max.
VSWR at I Port:	2.0:1 max.
(5—500 MHz)	

NOTES: 1. V_{cont} Set for I_o = 40 ma.
2. LO Level: −5 dBm typical.
* Measured At 10MHz

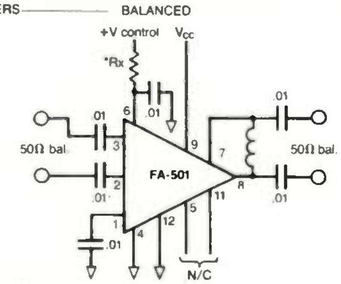
TYPICAL APPLICATIONS

AMPLIFIERS

UNBALANCED

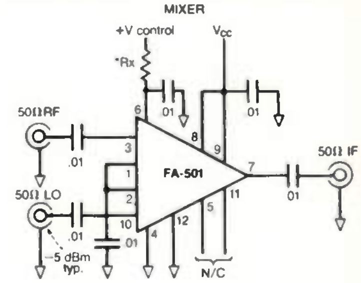


BALANCED



*R_x required for control voltages higher than +5 volts; R_x = 560Ω for $V_{CONTROL}$ = +15V.

MIXER



* Set $V_{CONTROL}/R_x$ for $I(V_{cc})$ = 40 ma.

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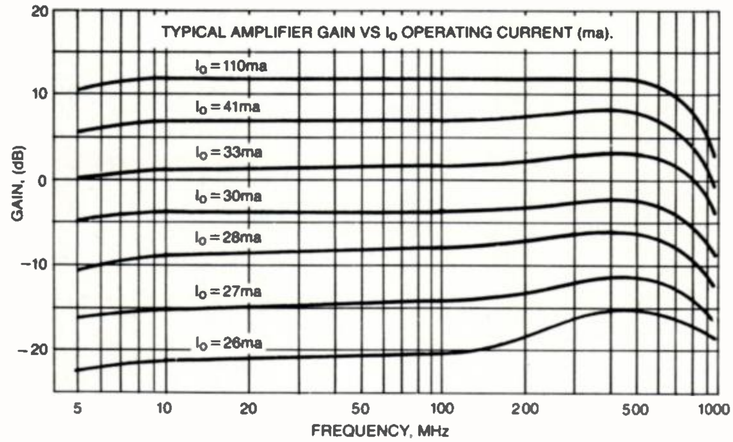
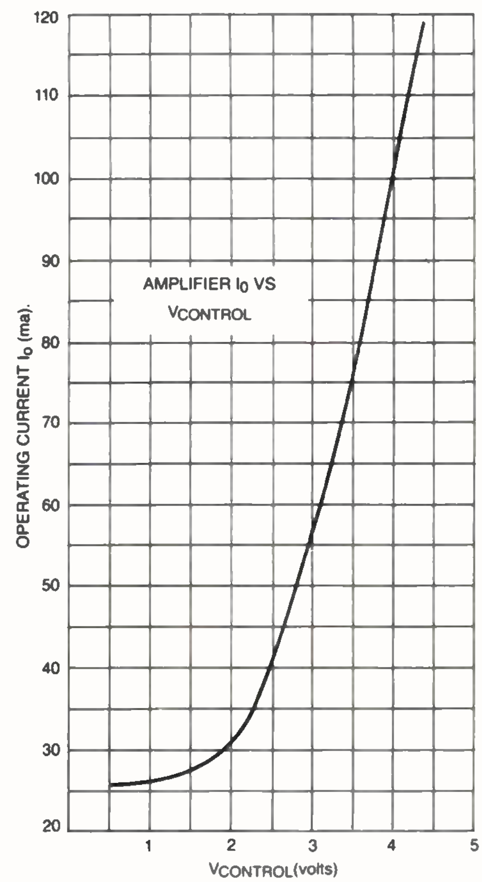
MICROWAVE MODULES & DEVICES, Inc.

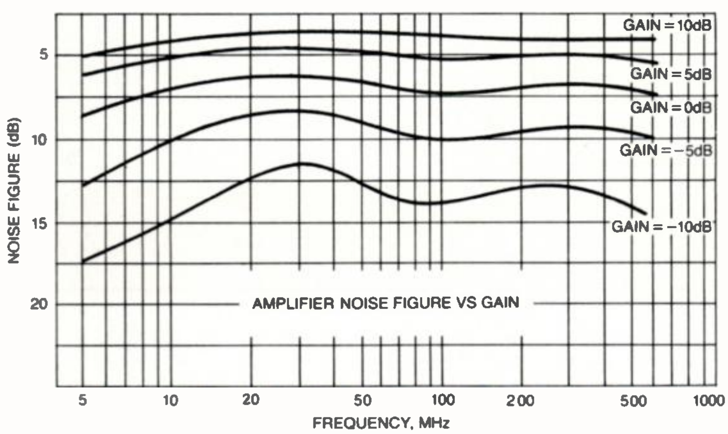
FA-501

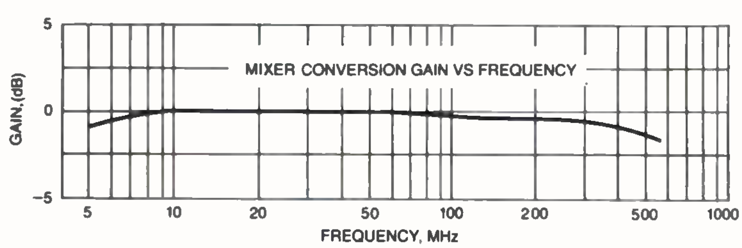
WIDEBAND RF FUNCTION AMPLIFIER

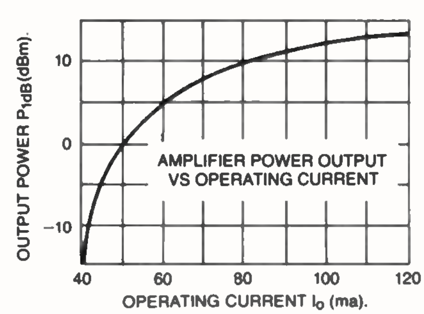
FREQUENCY RANGE: 5-500 MHz.

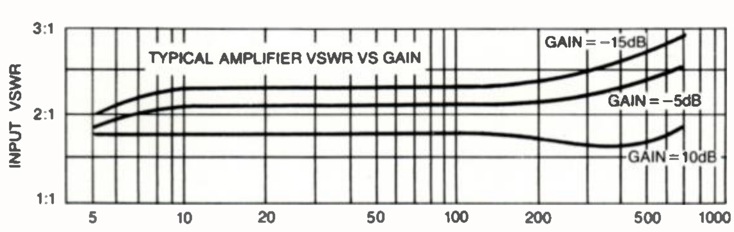


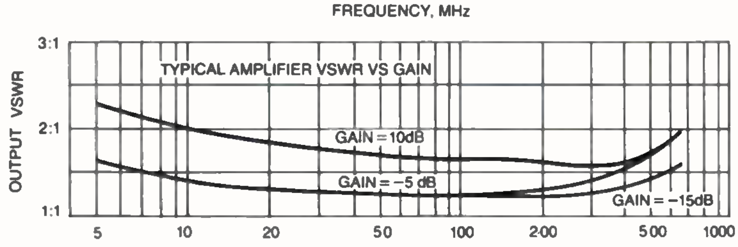


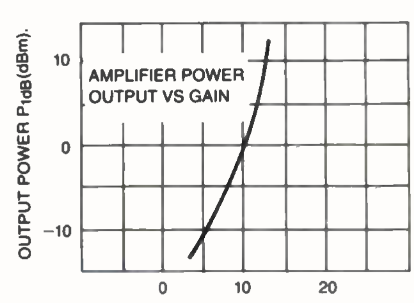












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AT $3950 IT'S A GIANT.

IT'S THE NEW 2022 • 1GHz SIGNAL GENERATOR

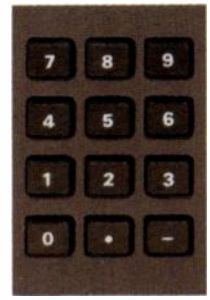
marconi instruments

marconi instruments

This 10 kHz – 1 GHz Genera

Main Function Keys

For entry of carrier frequency, modulation depth or deviation, and r.f. level. The associated l.e.d. is lit when a function is selected.



Numeric Keypad

Provides rapid and accurate setting of parameters.

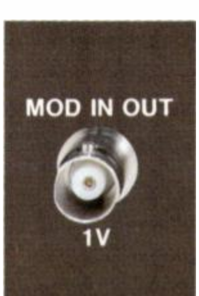
Δ Key

Allows step values of any size to be entered for the main functions.

The model 2022 weights only 16

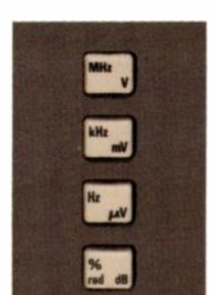
Store/Recall

Up to 100 settings may be stored, and recalled instantly. When Recall l.e.d. is lit the increment keys are used to step through memory locations in sequence, giving fast semi-automatic selection of settings.



Mod. IN-OUT

Provides wideband modulation input for external a.m., f.m. or ø.m. with selectable a.l.c. and out-of-limits indicators to ensure high accuracy. Also provides output of internal modulation.



Unit Keys

Terminate parameter entry. They also provide single-key conversion of r.f. level settings, e.g. µV to dBm.

' Sets New Price/Performance Standards

t/Ext
›lects internal or external fre- ıency standard, or toggles ›tween internal and external ɔdulation source.

Second Function
Gives access to many extra and innovative facilities. Sensitive Second Functions are protected by a secure key sequence.

; illustrated precisely full size!

Mod. ON-OFF
Removes modulation, for speedy simple signal-to-noise measurements.

Carr. ON-OFF
Toggles r.f. output on and off. For fast measurements of quieting sensitivity, etc.

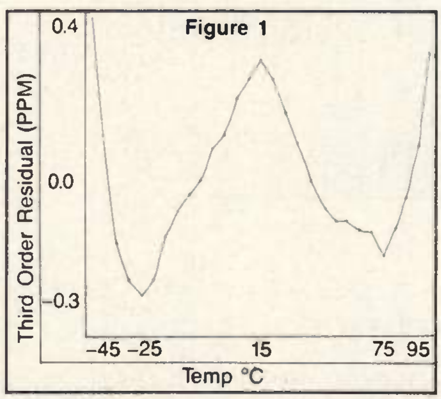
:rementing Facilities
ow any parameter to be remented or decremented in ›r-defined steps. TOTAL Δ plays the shift from initial :tings. RET'N key restores ial setting.

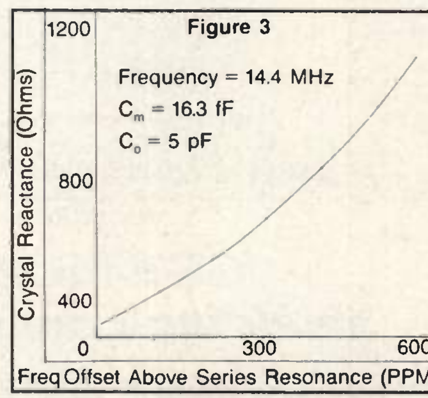
RF Output
Delivers −127 to +6 dBm in c.w. and f.m. modes. Reverse power protection up to 25 W saves accidental damage to output circuits. Protection fail-safe — acts even with unit turned off.

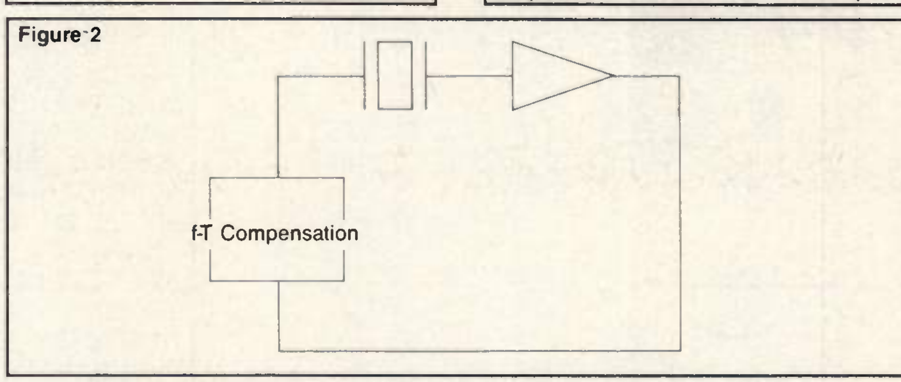
Frequency-Temperature Performance Determination In High Stability TCXOs

By Dennis Marvin
Motorola, Inc.
Communications Systems Division

The temperature compensated crystal oscillator (TCXO) has been with us for over thirty years and is quite often taken for granted. However, as frequency vs. temperature (f-T) stability requirements continue to tighten, many of the previously ignored second order effects will have a significant impact. TCXO stability requirements of 0.5PPM or 2PPM over temperature ranges as wide as –55 to +95°C are becoming commonplace. This article will review several of these second order effects and attempt to quantify the amount of performance debasement that can result.







AT-cut crystals are commonly assumed to have an f-T characteristic that can be represented by a third order polynomial. While this fit is quite adequate for most current applications, it can be inadequate for tight f-T requirements over wide temperature ranges. Figure 1 shows the difference between actual data and the least squares third order curve fit for a typical fundamental crystal. There is a decided, consistent fourth order component. Oscillators with f-T stabilities approaching 1PPM must incorporate a compensation scheme capable of correcting for the fourth order term and crystal evaluation algorithms must do a fourth order fit to insure that good crystals are not discarded.

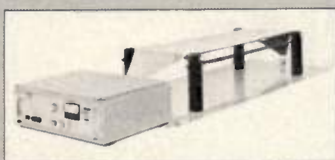
Tight tolerance TCXO designs are further complicated by component TC and non-linear transfer characteristics. Even though the frequency stability of an AT-cut crystal is almost always less than 1PPM/°C, work at Motorola and by Holbeche and Morley(1) has determined that the motional capacitance, C_m, of a fundamental crystal varies approximately 230PPM/°C. This variability must be addressed during the oscillator design. The varactor and its operating voltage must be carefully selected. If a varactor with a non-linear V-X characteristic is used, the compensation network will have to generate a more complex V-T function to compensate for the distorted varactor transfer function. Similarly, if a nonlinear varactor is used for both temperature compensation and aging adjust, the amount of compensation will change as the aging is corrected.

Background

A fundamental AT-cut (3-20 MHz) TCXO can be simply represented by the block diagram of Figure 2, where the feedback scheme is incorporated into the amplifier block. This block can represent any of the common configurations: Colpitts, Pierce or Clapp. The oscillator operates by meeting Barkhausen's criterion; the sum of the reactances around the loop must equal zero, with the crystal operating at whatever frequency meets this requirement. Figure 3 shows that the crystal reactance is a function of fre-

NEMP—Tester for EMP-Tests

The portable PEMI 12 tester simulates the fast impulses generated by high altitude nuclear explosions. Test your instruments and electronic assemblies with this tester delivering a<10/400 ns impulse with an adjustable peak voltage of 0.5 ->10 kV! The bus bar connector allows fitting a 150Ω parallel strip line antenna accommodating a standard 19" EUT and subjecting it to an



E-field of 50 kV/m. Separate current injection output for peak current up to 150 Amps!

If you care whether your equipment is NEMP-proof you should call us or send us the bingo card. We have a lot of experience with NEMP testers with extremely fast rise times, charging voltages up to 1.5 MV and we manufacture them also to NATO specs.

AMERICAN

HV TEST SYSTEMS, INC.

Telephone (301) 826-8651 Telex 86223
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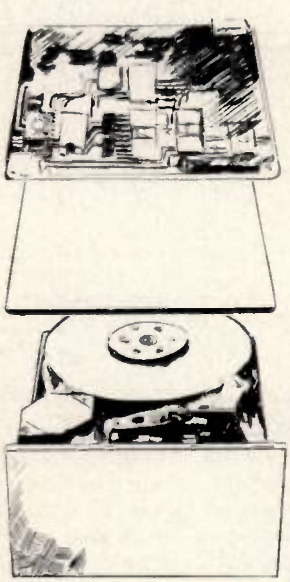
INFO/CARD 27

GST-MC2
Shielding Laminates



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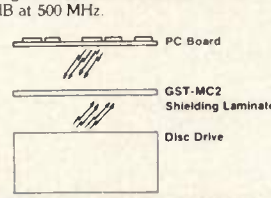

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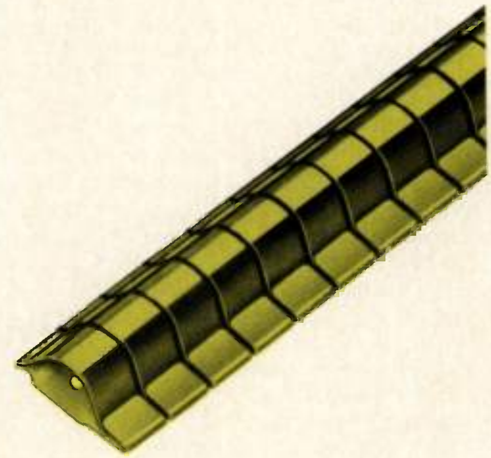


quency by following the nonlinear relationship:

$X = 1/((C_m/2\Delta f/f - C_o)w)$

Where: C_m is the motional capacitance of the crystal model;
C_o is the capacitance of the crystal plate, the holder and any associated stray circuit capacitance;
$\Delta f/f$ is the fractional frequency change of the crystal from series resonance.

Such a circuit can be temperature compensated by having a compensation block that generates a reactance characteristic as a function of temperature such that it causes the crystal to change the operating point on its reactance curve. By carefully designing this block the crystal can be operated so as to cause a net "zero" frequency change over temperature.

There are several networks that are used for the compensation block, both direct and indirect. The preferred circuits are indirect, i.e., a DC voltage is generated external to the oscillator loop and us used to vary the loop reactance via a varactor. Direct compensation, where the compensation is inserted directly in the oscillator loop, has fallen on disfavor, as its use lowers circuit Q. Of the several methods of indirect compensation used, including custom integrated circuits and digital compensation, the most common and the most easily understood is thermistor-resistor compensation. One embodiment of this is shown in Figure 4(2). Other designs can be similarly analyzed.

If we assume that the varactor has a $\partial = 1$ (not bad for hyperabrupt varactors over a limited voltage range) and that all other parts in the oscillator are constant over temperature ms. characteristic of the crystal. This is a good approximation for initial analysis purposes, with corrections necessary for an exact compensation.

In any case, this thermistor-resistor network has the following transfer function:

$V_{out} = V_{in} RT3(R1+RT1)\ (R2+RT2)/((R1+RT1)(R2+RT2)\ (R3+RT3)+R2RT2(R1+RT1+R3+RT3))$

where RT1, RT2 and RT3 are the resistances of the three thermistors at a particular temperature.

R1 and RT1 affect performance over the entire temperature range, while R2 and RT2 primarily affect the cold end and R3 and RT3 primarily affect the hot end. Care must be taken to obtain suitable parts values so the appropriate transfer function can be obtained while still maintaining a high degree of independence among the temperature ranges. The use of a personal computer to gather f-T data and process it reduces the need for independence among the compensation sections.

Parts Aging

When oscillators are being temperature compensated one is normally concerned with passing the specification at a half-dozen or so temperatures with a little margin for measurement error. This is quite adequate for oscillator stabilities of 5PPM of looser, but if a tighter system stability is required parts aging can become very important.

Parts aging can have two effects on the oscillator stability. The first is a positive or negative frequency shift over the entire temperature range, i.e., the oscillator will have the same f-T characteristic after aging has occurred as it did before, but the whole response has a new room tem-

TYPE	AGING
Resistors	
Carbon Comp	>10%
Carbon Film	3 to 5%
Thick Film	1 to 5%
Metal Film	1.5 to 2%
High Precision Metal	
Film Thermistors	0.1 to 0.5%
Non-hermetically sealed	1 to 5%

TABLE 1

perature reference frequency. A more insidious result of parts aging is the distortion done to the f-T response.

Over the life of a product, the resistors and thermistors can age as shown in Table 1.

The effects of parts aging on frequency drift and the f-T characteristic can be determined by inserting a small resistor or thermistor change (e.g., 1%) into the transfer function and analyzing the results. Parts changes of + and - 1% for all of the resistors and thermistors in Figure 4 were inserted and the resulting errors maximized by proper selection of sign. The maximum offset was 36mV or 0.61PPM given a sensitivity of 17PPM/V. Figure 5 is a plot of the maximum f-T distortion. Changes in f-T performance as poor as 0.80PPM/% change are possible. Allowing the 1% aging to occur in a random direction among the resistors and thermistors, there is on average a 0.27PPM/% offset and a 0.41PPM/% f-T distortion to the T.C. curve.

Of these figures the f-T distortion is by far the more serious problem, since a means of frequency adjust can usually be provided to compensate for the frequency drift. With these results now available, the appropriate resistors and thermistors can be specified and the initial f-T specification appropriately tightened so that system stability can be maintained to allow for compensation aging.

Frequency Adjust Aging

Compensation aging isn't the whole story, unfortunately. The major source of frequency drift is almost always the crystal. This drift is normally corrected by adjusting a capacitor or inductor in the oscillator loop to shift the operating point in the crystal curve and thus return the oscillator to nominal frequency. This shift in operating point also distorts the f-T characteristic and is known as trim effect. Galla and McVey(3) have developed equations to quantify this distortion for both series and parallel adjust circuits (see Figure 6). The values shown are typical for a fundamental frequency equivalent circuit. These equations are:

f-T distortion (PPM)

SERIES

$$\frac{4apC_o(C_o+C1)X10^{-3}}{C1C_m}$$

PARALLEL

$$\frac{4ap(C_o+C1)\times10^{-3}}{Cm}$$

where a is the amount of aging to be corrected (PPM)
p is the amount of temperature compensation (PPM)
C_o is the crystal and holder capacitance (pF)
C1 is the load capacitance seen by the crystal (pF)
C_m is the motional capacitance of the crystal (fF).

As the distortion in the series adjust configuration is equal to that of the parallel adjust multiplied by C_o/C1, and this term is significantly less than 1, the series performance far surpasses that of the parallel. The parallel adjust should never be used in a high stability design if the use of a series is possible. Even if a series adjust is used, an additional 0.17PPM distortion will be generated. Figure 7 shows the f-T characteristic for a series adjust oscillator and its trim effect.

There are several ways that this problem could be alleviated. If C_m were increased, the distortion would decrease. However, all things being equal, C_o would increase in proportion to C_m. Additionally, the larger C_m becomes, the more the variations in the oscillator components affect performance.

The easiest way to reduce this distortion is to reduce p, the amount of temperature compensation, by carefully selecting the crystal angle. It can be further reduced by not overspecifying the temperature range. The smaller the temperature range, the smaller p will be.

Reducing a, the amount of aging to be corrected, is also possible. This can be done by using a better aging design, preaging the crystals or by sorting them for aging. Remember that a reasonably well made crystal will age approximately as much in the first year as it will age in the next ten years combined.

Voltage Regulator Contributing

In addition to the compensation circuitry and the crystal aging, one must also consider the voltage regulator circuit driving the compensation. The regulator is very important, as any voltage variations in it are directly coupled to the varactor and will create drift and distortion problems.

Using the compensation network of Figure 4 and allowing the regulator to vary +1mV/° (a fairly typical value for a fixed commercial regulator) results in an f-T distortion of about 0.4PPM, as shown in

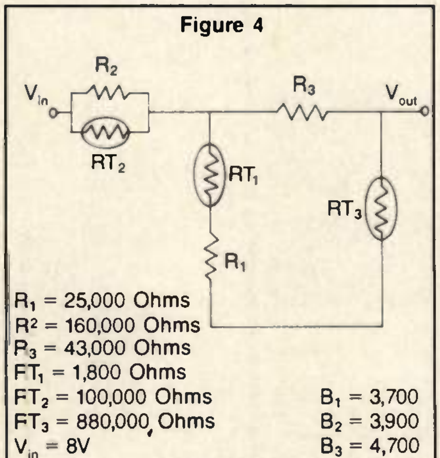


Figure 4

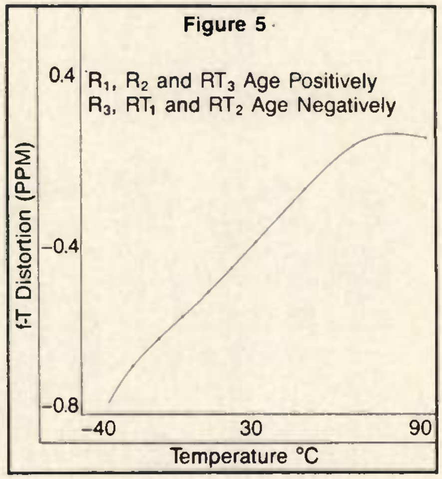


Figure 5

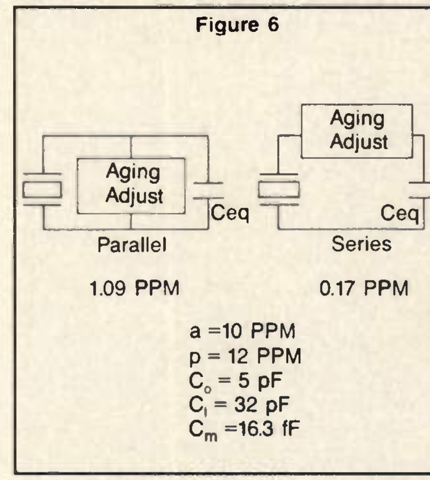


Figure 6

Figure 8. The only way to alleviate the problem of regulator voltage variation with temperature is to temperature compensate the oscillator with an internal regulator so that the temperature variations are compensated out.

Even with a built-in regulator, there is still the potential for aging. Figure 9 shows the distortion for a +1% change in regulator voltage.

Hysteresis

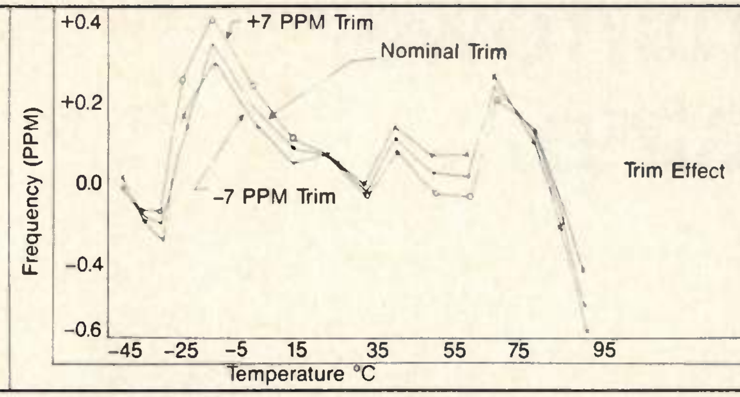
Crystal oscillator frequency performance can be very dependent on environmental dynamics. The designer must consider what dynamic frequency stability is required and what maximum rate of change of temperature the oscillator will see. The temperature of the oscillator does not normally change nearly as fast as the ambient, as it is generally well isolated inside the system.

There are two major contributors to the hysteresis problem. The most easily visualized is the imbalance in thermal time constants among the several thermistors and the crystal. Time constants for thermistors in still air range from less than 0.1 sec. for small chips to over 10 sec. for 0.1 inch diameter beads, while typical HC-18 crystals have time constants of from 1 to 2 minutes. There is a lot of variability in the crystal time constant. Important parameters causing this variability include package, lead and mount material and size, crystal blank size, crystal mounting technique and the type and density of gas sealed in the package.

There are so many variables present that the best way to determine the magnitude of the frequency shift is to measure it for the particular application. Tests performed on a hand-held 2-way radio subjected to a −30 to +90° temperature shock have given shifts as large as 0.3PPM.

If a product will be subjected to severe

Figure 7



swings, this effect can be reduced by better matching time constants through thermally coupling the compensation temperature sensors to the crystal or by increasing the shorter time constant.

The second hysteresis problem lies within the crystal. If an AT-cut crystal undergoes a temperature change, stresses will be set up within it. As the temperature of the crystal is swept back and forth an f-T characteristic is generated that traces a loop about the familiar static characteristic. The direction and magnitude of this offset is dependent upon the particular crystal design, the temperature range and the speed with which the temperature is changed.

Ballato(4) has modified this effect by changing the familiar crystal model from

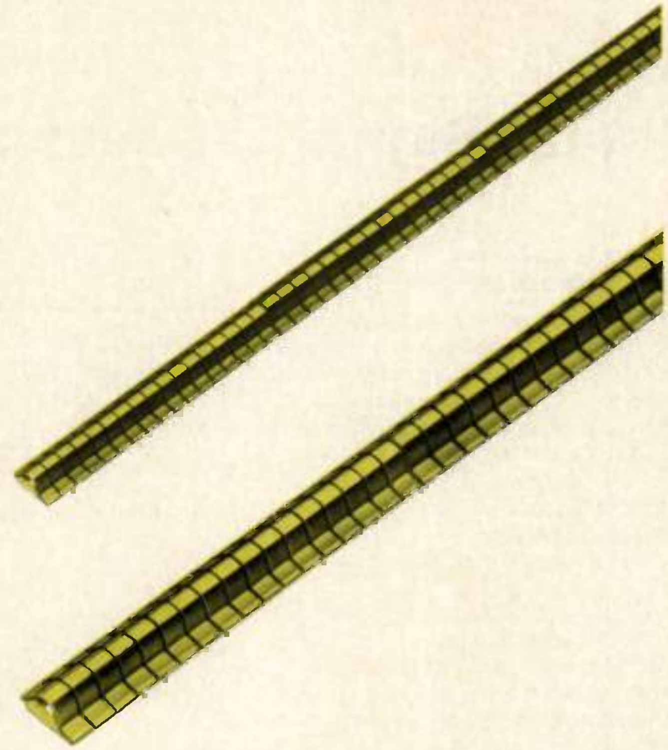
$$f/f = a_0\Delta T + b_0\Delta T^2 + c_0\Delta T^3$$

to

$$f(t)/f = a(t)\Delta T(t) + b_0\Delta T^2(t) + c_0\Delta T^3(t)$$

where $a(t) = a_0 = \hat{a}T(t)$ with $\hat{a}$ being evaluated separately for each crystal design.

For a particular design where the



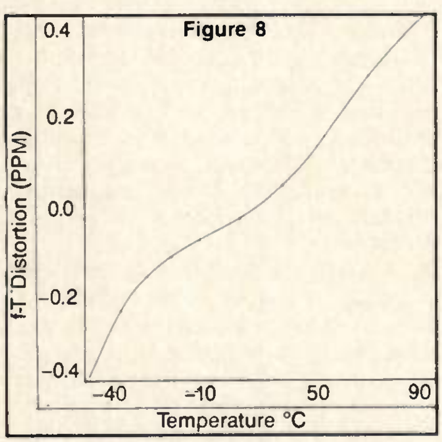


Figure 8

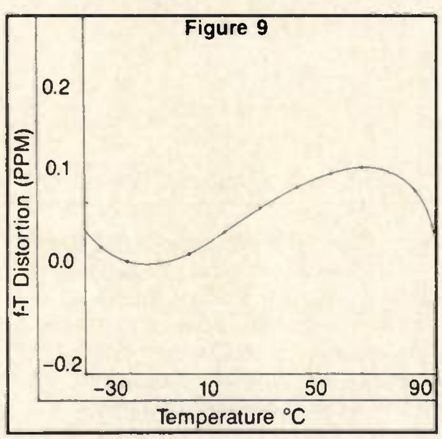


Figure 9

temperature is varied in a sinusoidal manner + and - 60° about 30° at a frequency of 0.01Hz (an orbital time of 628 sec.) the peak deviation from the static curve was 7.2PPM. As the deviation is linearly related to the frequency of the temperature change for this model, the maximum orbital time to keep the peak deviation below 0.1PPM for this crystal is 12.6 hours. Hence, fast temperature changed can and do affect the desired stability. It should be stressed that the magnitude and direction of this deviation is quite crystal dependent, with this example tending to be worst case, and that this is the peak error over the entire temperature loop.

Conclusion

High stability TCXOs are subject to many instabilities not normally considered in the design of lower stability products. Several of these have been reviewed. The design engineer must carefully consider the system requirements in dynamic environments and as a function of time.

For example, the design Motorola has chosen to use is shown in Figure 10. The as shipped f-T stability requirement is as tight as 1PPM from -40 to +95° with a total system stability of better than 2PPM. In order to accomplish these specification and still maintain a small package size (< 0.3 cu.in.) while minimizing costs, a custom integrated cicuit in a hermetic chip carrier was used (5). The chip consists of a three segment non-linear compensation ms, but has improved isolation between the stages for ease of production. Compensation is achieved with four trimmable resistors. Hermetic sealing of the IC minimizes regulator drift. Parts drift s minimized by utilizing the matched TC characteristics in the IC design while good crystal aging is maintained by preaging and then sorting crystals with a curve fit routine using data obtained by actively aging the crystals. Hysteresis is also minimized through crystal sorting and with attention to thermal design. Overall hysteresis and aging are further controlled by hermetically sealing the entire oscillator. rf

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Digital Signal Processing

By Thomas Callaghan
Watkins-Johnson Company

Digital signal processing is not an easy topic to comprehend. It combines theories of both the analog and the digital disciplines. To ease this understanding, one should remember that all aspects of DSP can be divided into three areas:

1) Input Signal
2) Digital Processor
3) Information Extraction

The input signal must be sampled and quantized before it can be fed to the digital processor. The digital processor can only perform the operations of addition, multiplication and delay (storage) on the signal data. The end result of the digital processor is to extract information from the signal or modify it in some way. This article will explain the input signal.

The input signal, which is analog in nature, can be classified as a continuous time, continuous amplitude signal. To be of a form useful to a digital procesor the signal must be converted to a discrete time, discrete amplitude signal.

Sampling

Sampling is the process by which the continuous time, continuous amplitude signal is converted to a discrete time, continuous amplitude signal. This is done by periodically taking minute time chunks out of the analog signal. The output of a sampler is a series of pulses whose envelope approximates the original signal. The process can be represented by multiplying the input signal by a pulse train of uniform amplitude and equal spacing, (Figure 1).

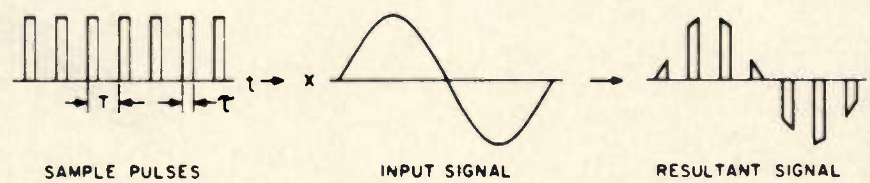

Figure 1. Characteristic Signals in Sampling

The pulse spacing, T, that is the sampling frequency, 1/T, cannot be arbitrarily chosen. According to the Sampling Theorem, "If a continuous time function contains only frequency components below F cycles per second, 2F samples per second suffice to represent it perfectly and permit perfect recovery." The reason for this is more readily seen in the frequency domain. Referring to Figure 2, one can see that one of the byproducts of sampling is the duplication of the signal every F_s, the sample frequency. If the sample frequency is kept above or equal to 2F, (Figure 2b), the sampled frequency spectrum produces no overlap and the original signal can be retrieved. If the sample frequency drops below 2F, the sample frequency spectrum starts to overlap. Any of the frequencies whose spectra are contained in this overlap tend to fold over about $F_s/2$ and into the actual frequency band, as indicated in Figure 2C. This foldover of frequency is called aliasing.

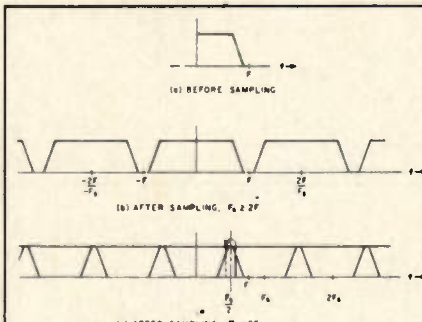

Figure 2. Frequency Domain

Examples of frequency aliasing are witnessed daily by many unknowing observers. For example, in the old "Westerns," the stage coach rides out of town and its wheels appear to speed up, stop and then go backwards. The backwards revolution is caused by the movie camera not taking sufficient frames per second to dependably capture the wheel's revolution. Another example is when a television or computer terminal is shown in a movie with black bars across the screen. Again, the movie camera is not updating the screen fast enough. In this case, the camera moves at 24 frames per second while the television scans a new frame at 60 times per second.

The sampling theory is all well and good on paper, but in reality it is difficult to remove all unwanted frequencies above the frequency of interest. The most common solution is to place a filter ahead of the sampler to remove unwanted frequencies. The filter, however, has two disadvantages; it alters the signal of interest and unless it has very steep skirts certain unwanted frequencies are passed. A second solution is to sample at a faster rate than 2F. How much faster is open to debate. Sample too fast and one puts an undue burden on the digital processor.

Another consideration in determining sampling frequency is interpolation errors. Since one is dealing in discrete time, what has happened to the input signal between samples is not known. Naturally, the faster the sample rate the more information known about the signal. When processing is finished, it may be desired to convert back to an analog signal. The conversion is usually done with a D/A converter and some type of low-pass filter to smooth, or interpolate, the discrete points.

So how does one select the sample rate? Gardenhire in [1] presents a good approach to the problem. In his work, the sampling rate is based upon the amount of interpolation error tolerable from the output filter given a specified input filter roll-off characteristic. The input filter roll-off determines the order of the system. Given a certain roll-off, m, the sample rate can be determined from the acceptable interpolation error. The rate is given for a 5% error in Table 1.

The filter method employs analog interpolation after D/A conversion. The 2, 3, or 4 point linear interpolation method is done via a computer and the output is still in a digital format. It is interesting to note how the sample rate approaches the theoretical rate of 2F as the order of the filter increases. It is also worth noting how the sample rate increases drastically as one tries to follow the input signal more closely with a minimum of error.

For communications applications, however, one is dealing primarily with input filters whose order does approach infinity. The output filters are also optimized, or of Butterworth types. For these reasons a sample rate in the range of 2 to 5 times F should prove satisfactory.

Another error that appears during sampling is known as aperture error. This error is caused by the sample pulses taking a small amount of time to capture the analog signal. During this time, the input voltage can change drastically as the signal goes through zero. This is indicated in Figure 1 by the triangular shapes of some of the sampled data pulses. The aperture error can be expressed mathematically [2] for a sinusoidal input by:

$$\varepsilon = \frac{\Delta V}{V_{FS}} = 2\pi\tau,$$

Table 1. Systems Resulting in a 5% Interpolation Error [1]

Normalized sample frequency (F_s/f_1)

Interpolation Method	m = 1	m = 2	m = 3	m = 4	m = 5	m = 6
Wiener Optimum Filter	640	11	5.1	3.8	2.6	2.0
Butterworth n = 4	—	16	8.3	5.5	5.5	5.5
n = 3	—	18	9.2	6.7	6.7	6.7
n = 2	1.2×10^3	29	17.0	11	11	11
RC Filter n = 1	1.2×10^4	220	130	91	91	91
2 Point and Linear Interpolation	640	13	8.3	5.9	5.9	5.9
3 Point Linear Interp.	640	12	6.2	5.2	—	4.0
4 Point Linear Interp.	640	12	5.7	4.3	—	3.3

where V_{FS} is the full scale voltage and τ is the duration of the pulse.

Expressed another way, the acceptable pulse duration can be calculated as

$$\tau = (2\pi \times F \times 2^n)^{-1}$$

where n is the number of bits of resolution of the A/D converter for an A/D converter with a 1LSB error. For instance, if the input frequency is 10 kHz and eight bits or resolution are used, the allowable pulse duration = 63.5 ns.

To alleviate aperture error, a track and hold amplifier can be inserted before the A/D converter. This device's output will follow the input signal while its track digital level is active. When the track input switches, the output is held at its current level and will not vary over the duration of the pulse.

Quantization

After the discrete time, continuous amplitude signal is obtained, it must then be converted to a discrete time, discrete amplitude signal for the digital processor. This process is formally called quantization, but it is more commonly known as analog to digital (A/D) conversion. A quantizer takes a specific amplitude range and divides it into a series of discrete steps, Q. A digital number is then assigned to each Q.

The number of bits in the digital word determines the number of steps that can be achieved. For n bits, the number of steps would be 2^n. Each step Q, or each change of 1LSB, would be

$$Q = V_{FS}/2^n$$

where V_{FS} is the full scale magnitude of the allowable input voltage.

If the input signal falls between steps then the digital number assigned to it depends on whether the quantizer rounds the samples or truncates them.

As its name implies, the quantizer rounds off the analog input to the nearest quantizer level, Q. In truncation, (Figure 3), the signal is represented by the highest Q level that is not greater than the signal. Thus, for each LSB (least significant bit) step of the digital output word, the error from the original signal would be $\pm 1/2$ LSB for rounding and a 0 or +1LSB for truncation. An error of $\pm 1/2$ LSB yields a mean error of zero, whereas a 1LSB error yields a mean error of $+1/2$ LSB. For this reason, rounding is preferred in most practical considerations.

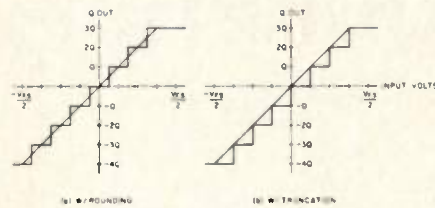

Figure 3. Quantizer Characteristics

How the digital word is represented is another area of consideration. Binary representation varies widely for positive and negative numbers. Again, for most practical considerations, two's complement representation is chosen because most processors use this type of representation. Also, for other applications only positive numbers are expected. Many A/D converters allow either; that is, a range of 0 to V_{FS} or $-V_{FS}/2$ to $+V_{FS}/2$, as well as other representations.

Finally, once the quantizer or A/D converter is chosen the number of bits of resolution must be decided. While a large number of bits will represent an analog signal more accurately, they will not represent a cleaner analog signal. In every analog signal, there exists some inherent noise that is some small portion of that signal. The more bits of resolution, the smaller the step size in the quantizer. The smaller the step, the more the digital word is affected by noise, so that the lower significant bits only serve to give a good representation of this noise.

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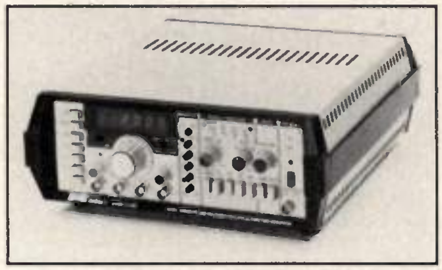
Function Generator

The Model 625 5 MHz Function Generator offers sine, square, triangle, haver waveforms, inverted waveforms, variable symmetry waveforms and DC. Frequencies from 100 microhertz to 5 MHz may be selected using 9 overlapping ranges. Within each range except 5 MHz the frequency may be varied over a ratio of 1300:1 remotely, from the front panel or using an external VCF voltage. In the 5 MHz range the frequency may be varied over a ratio of 500:1. Operating modes include: Run (continuous waveform), Gate (continuous only while gated), Trigger (single cycle waveform with each trigger) and Inhibit (no waveform at output except during GPIB trigger commands. i.e., external and manual triggers are inhibited).



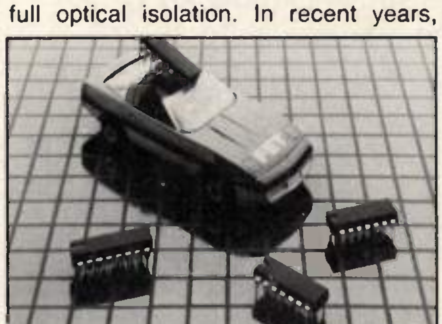
All numerical parameter values are selected and displayed on a 3½ digit LED display. These parameters are: frequency, amplitude, offset, symmetry and calibration sequence number. A decimal point and range LEDs determine the multiplier. Amplitude range is 10 millivolts p-p to 10 volts p-p into 50 ohms, offset range is zero volts to ±5 volts peak into 50 ohms, and symmetry range is 5%:95% to 95%:5% with 1% resolution. Additional front panel LEDs annunciate the non-numerical parameters which are selected. **Exact Electronics, Carson City, Nev., INFO/CARD #176.**

Hyperabrupt Tuning Varactors

DKV4105 A, B & G and DKV4109 A, B & G hyperabrupt tuning diodes are packaged in glass rather than plastic to offer hermetically sealed devices for all applications. The high reproducibility of C vs. V characteristics guarantees long-term availability of devices with uniform electrical properties. The diodes are ideal for straight line frequency vs. voltage applications in crystal or LC-tuned circuits as well as for frequency or phase modulators. **Alpha Industries, Woburn, Mass., INFO/CARD #175.**

Isolated Power MOSFET Driver

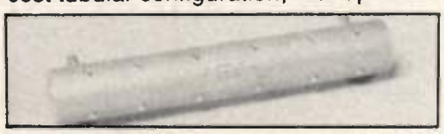
A 16-pin DIP, power MOSFET driver, the FDA100 provides 2500 volt input/output isolation. Now you can drive power MOSFETs in 10 to 500 microseconds with full optical isolation. In recent years,



power MOSFETs have gained increasing acceptance, especially in switching power supply applications. In addition to expected major use in programmable controllers, it can just as easily be used in high speed driving of smaller MOSFETs. Linked, for example, to a small signal FET, it can result in a speed/gain product for high speed datacom circuits much greater than that achievable with opto-coupler circuits. **Theta-J Corp., Wakefield, Mass., INFO/CARD #174.**

Tunable Bandpass Filters

A full line of n=2 > 12 pole (easy cavity adjust) bandpass filters, in convenient low cost tubular configuration, with up to 5:1



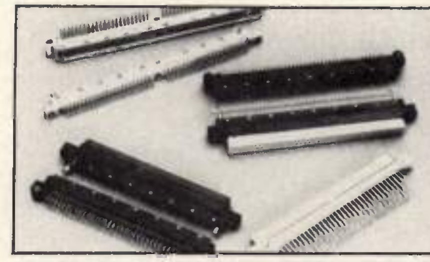
tuning range (example: 0.2 > 1 GHz) is now available. **Cir-Q-Tel, Kensington, Maryland, INFO/CARD #173.**

MATE Interfaces

MATE Interfaces for the 50 MHz Pulse Generator Model 859 and the 50 MHz Synthesizer Model 178 have proven records in systems based on GPIB (IEEE-488) interface. The new MATE interface allows GPIB programming in user friendly CIIL (Control Interface Immediate Language) which standardizes intrument command syntax. In both models, a full MATE confidence test capability is implemented which assures the user that the unit functions properly. In addition to the MATE interface both models also feature the standard GPIB interface. **Wavetek, San Diego, Calif., please circle INFO/CARD #172.**

PX Connectors

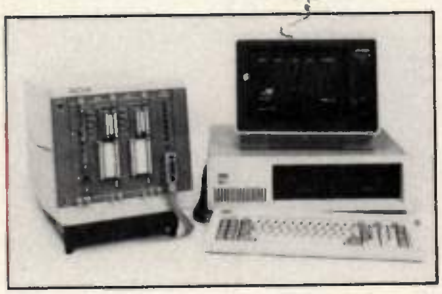
Printed circuit board connectors, featuring low insertion force and superior operation in rigorous environments of temperature, vibration and moisture are intermountable with and designed to meet or exceed MIL-C-55302 function requirements. The PX connectors provide an 80% reduction in connector mating force — 1½ ounces per contact. High performance thermoplastic insulators provide high strength at a service temperature range of –55°C to +125°C.



PX connector pins are shrouded to guard against field damage, and a polarized insulator housing prevents 180° mismating of the connectors. The connector design deals against moisture that can contribute to circuit or system malfunction. **ITT Cannon, Fountain Valley, Calif., please circle INFO/CARD #171.**

IMB PC Based Multichannel Digital Oscilloscope

WAVEFORM-CATALYST is a multichannel digital oscilloscope software package with powerful capabilities for acquisition, display, archiving and analysis of high-speed waveforms. It is an integrated system consisting of LeCroy's series of modular waveform digitizers, any of the IBM personal computers (PC, portable, XT or AT) and system software. Digitizer plug-ins offer bandwidths to 100 MHz and maximum sampling rates of 200 MHz or 5 nsec between points. CATALYST operates like an instrument, not a computer. Single keystroke operating commands and familiar scope terminology make knowledge of computers or programming language unnecessary. The high resolution graphics permit simultaneous viewing of up to four



"live" or previously stored traces, each with its own timebase and grid. Dual cursors control zoom, positioning, and the absolute and differential measurement of time and voltage, while on-screen markers identify the trigger point, zero baseline, off-scale conditions, etc. **LeCroy Research Systems Corporation, Spring Valley, New York, please circle INFO/CARD #170.**

HP Oscilloscope

The HP 54200A/D is a fully programmable, 200-megasample-per-second digitizing oscilloscope, ideal for capturing high-speed single-shot events. It is particularly useful to designers of digital and analog systems who must deal with transient fault conditions. Transients as narrow as 10 nanoseconds can be captured easily. The HP 54200A/D's 200-megasample-per-second digitizing rate enables it to capture high speed events in a single-shot manner. Pre-trigger viewing allows the user to see what happened before the trigger event. With built-in automatic pulse-parameter measurements, the HP 54200A/D provides fast, repeatable

answers. Automatic answers relieve the designer of having to write programs to cover these functions. The HP 54200 A/D's average mode extracts the signal from uncorrelated noise, allowing the user the probe in noisy environments. The HP 54300A/D's infinite persistance eliminates the trial-and-error fine-tuning inherent in analog storage controls. With two proprietary analog-to-digital converters, the HP 54200A/D provides two-channel simultaneous acquisition without sacrificing sample rate. The HP 54200A/D provides four memory locations in which a waveform, averaged waveform or waveform envelope can be stored for later analysis, allowing the user to make before and after comparisons. **Hewlett-Packard, Palo Alto, Calif., INFO/CARD #169.**

Coax Dual Line Strippers

The TOR-1C Coaxial Cable Stripper and the TOR-1F Dual Line Stripper are an ideal complement to the electrical and electronic technicians tool kit. Each tool comes with a spare set of blades. **Davle**

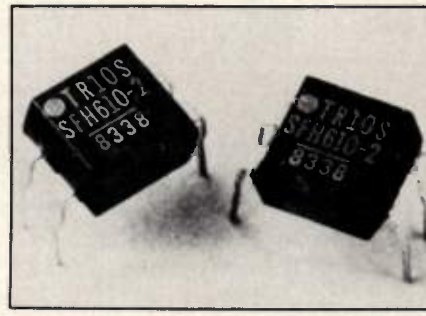
Tech Inc., Fair Lawn, New Jersey, INFO/CARD #168.

PIN Diode Drivers

Model 254's small 1/4 wide x 3/8 long size makes it easy to fit into any circuit. This inverting driver with typical output current of 20 mA is fully TTL compatible, including low power Shottky logics. It operates from ±5 VDC and dissipates less than ½ watt under switching conditions of 35 ns max. In house screenings per MIL-STD-883C, Method of 5008.2, Class B, Table VIII is available for a minimal charge of $10 each. **New England Microwave Corporation, Hudson, New Hampshire, INFO/CARD #167.**

Miniature Optocouplers

The SFH 610/611 series of miniature optocouplers in a 4-lead dual in-line package is being offered in four high current transfer ratio groups: SFH 610/611-1: 40-80%; SFH 610/611-2: 63-125%; SFH 610/611-3: 100-200%; and SFH 610/611-4: 160-320%. The difference in the SFH 610 and SFH 611 is only in the pin connections. Package dimensions for this handy 4-pin single-channel coupler are significantly scaled down in length to only .2" — compared to a standard 6-pin DIP length of .34" — and make it especially suitable for applications where board space is at a premium. It is designed for use in electronic test equipment, medical instrumentation, computer



peripherals and communications equipment. The devices are optically coupled isolators employing a gallium arsenide (GaAs) infrared emitter and a silicon phototransistor detector. They feature a 2800 volt isolation and are awaiting VDE approval. **Siemens Components, Inc., Cupertino, Calif., INFO/CARD #165.**

TWT Power Supply

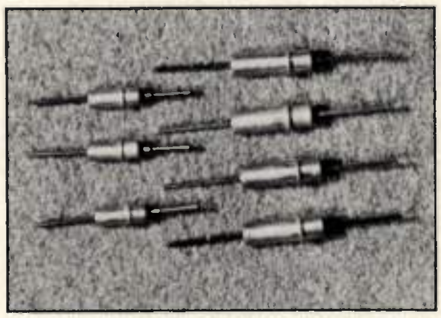
A MODULAR INTEGRATED TWT POWER SUPPLY, Model No. 10755 coupled to a 50-125w (C-Band) or 50-125w (Ku-Band) TWT tube will reduce waveguide losses for Ku-Band and C-Band antenna pedestal installation. Highly efficient with state of the art current driven VFET circuitry, the power supply is easily adaptable to particular customer specifications and unique requirements. Fully encapsulated high voltage components are included in this easy to operate unit with capacity up to 125 watt TWTs. The power supply has a modular concept, extreme wide operating temperature range and is conduction cooled. **MCL, Inc., LaGrange, Ill., INFO/CARD #163.**

Circuit Simulator

The SIMON circuit simulator is a full circuit simulation software package created specifically for MOS and CMOS circuits. Using an entirely new algorithm (not a SPICE derivative) SIMON offers unmatched speed (10 to 50 times faster than SPICE), guaranteed convergence, simulation of circuits far larger than previously possible and state of the art user friendly features. SIMON is available for a variety of host computers with price depending on the individual installation and CPU unit. Ongoing support, maintenance and updates are available via a maintenance agreement. **Simon Software, Inc., San Jose, Calif., INFO/CARD #162.**

Ruggedized EMI Filter Pins

EMI Filter Pins are improved and updated versions of the traditional tubular ceramic-bodied filter pin. Features include: metal shell construction providing a rugged housing to withstand assembly and environmental rigors; round or square cross section pins for solder or wire-wrap applications; expanded capacitance range. 100 pf to .07 μf available; voltage ratings of 50-300 VDC and 115 VAC; current rating 10 AMPS; available in C, L, π and T circuit configurations; typical insertion loss values of 4 dB at 100 kHz, 20 dB at 1 MHz, 50 dB at 10 MHz and 70 dB at 100 MHz and up for the 50 volt rated π section filter. **Ceramic Devices, San Diego, Calif., INFO/CARD #161.**



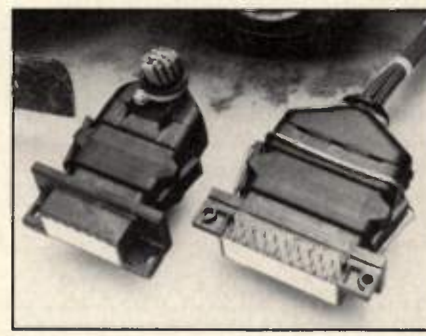
Digital Potentiometer

The HEDS-7500 digital potentiometer produces digital output directly from manual rotary input. The new device

enables bi-directional counting by decoding two digital-pulse trains, which are out of phase by 90 electrical degrees. Available with a standard 256 pulses per revolution, the device commonly is used as a knob on the front panel of equipment where a digital display may be used to indicate the position of the knob. **Hewlett-Packard, Palo Alto, Calif., please circle INFO/CARD #160.**

Snap Lock Connectors

An environmentally sealed series of quick connect/disconnect connectors that provides a labor-saving alternative to traditional environmental PC board connectors, called Snap Lock Environmental (SLE), is specifically designed for computer/controlled applications that must



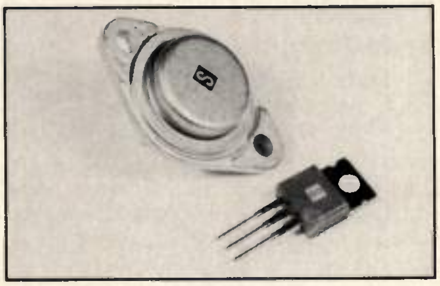
withstand severe environments. The SLE design does away with expensive slide latches, spring latches or even center jack screws by integrally molding snap locks to the receptacle body on the mating side. This side also offers a heavy duty interfacial gasket and multiple labyrinth seals in the rear to protect against severe conditions. **ITT Cannon, Fountain Valley, Calif., INFO/CARD #159.**

Insulated Ground Straps

CHO-STRAP™ Grounding Straps are flexible, double-insulated laminates of copper (inside) and Mylar (outside), with pre-punched tinned ends to facilitate termination. They provide low-impedance ground paths at RF frequencies, thereby reducing radiated EMI problems caused by ordinary wire grounds. CHO-STRAPS are generally more cost-effective than braided ground straps. CHO-STRAP ground straps are supplied in a variety of standard widths and lengths. Standard construction is 5-oz. rolled annealed copper, laminated on both sides to 2-mil Mylar with flame retardant polyester adhesive. CHO-STRAPS are UL rated and pass all requirements of UL specification 94VO. **Chomerics, Woburn, Mass., please circle INFO/CARD #158.**

800 Volt Hi-Rel Power MOSFETs

A new series of high reliability, fast switching power MOSFETs meet high temperature requirements and feature hexagonal geometry, voltage (drain to source) to 800 volts and drain currents to 456 amps. Identified as SWITCHMOS,



these units offer eutectic mounted packaging for military and industrial applications in TO-220 and TO-3 cases. **Solitron Devices, Inc., Riviera Beach, Fla., INFO/CARD #157.**

Miniature Pin Switch

This miniature pin switch has a frequency range of 10-1000 MHz at a 50 ohm impedance with an environmental temperature range of –55°C to +85°C. It

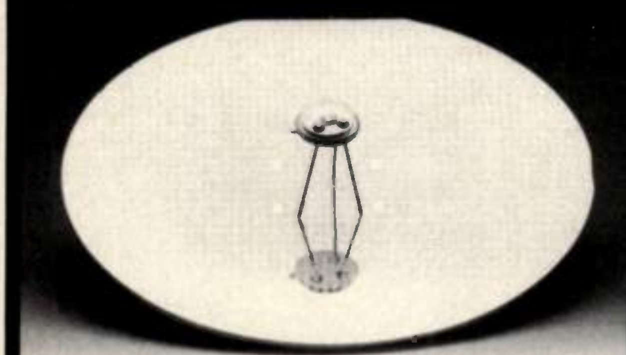
rf products Continued

is available in the TO-5 housing. Power requirement is ±15 volts at 10 mA, and a maximum RF input at 20 dBm. This SPST configuration has an insertion loss of less than 1.75 dB at 1000 MHz and isolation greater than 30 dB at 1000 MHz. **Wavetek Indiana, Inc., Beech Grove, Ind., INFO/CARD #156.**

Modular Probe Family

A unique high voltage RF detector probe, Model 10851A, features an optional "Tee" connector for direct 50-ohm measurements and a 100:1 divider to handle very high RF input voltages; the first multimeter probe capable of this performance. The new probes are designed to mate with virtually any brand of laboratory-grade oscilloscope. An adjustable wideband probe cable terminator network is provided so that correct high frequency compensation can be obtained for any scope. There is no compromise in risetime or bandwidth relative to comparable original equipment probes. Switch-selectable X1/X10 attenuation is also available on several models. **Ballantine Laboratories, Inc., Boonton, New Jersey, INFO/CARD #155.**

A quartz resonator for UHF oscillators has surfaced...



Introducing the surface acoustic wave (SAW) resonator. This fundamental-mode UHF resonator is manufactured with semiconductor processing techniques using quartz as a substrate. The high-Q resonance of the SAW resonator arises from phased reflections across the surface of the device — similar mathematically to the resonance of a laser. The quartz SAW resonator makes an excellent frequency control device for UHF oscillators from 300 to over 1000 MHz. SAW resonators are used in RF applications from precision instrumentation to high-volume consumer electronics.

For more information on the characteristics and applications of the quartz SAW resonator please contact:

RFM

RFMonolithics, Inc.
4441 Sigma Road
Dallas, Texas 75244
(214) 233-2903
TWX: 910-860-5474

rf literature

Electrolytic Capacitor Catalog

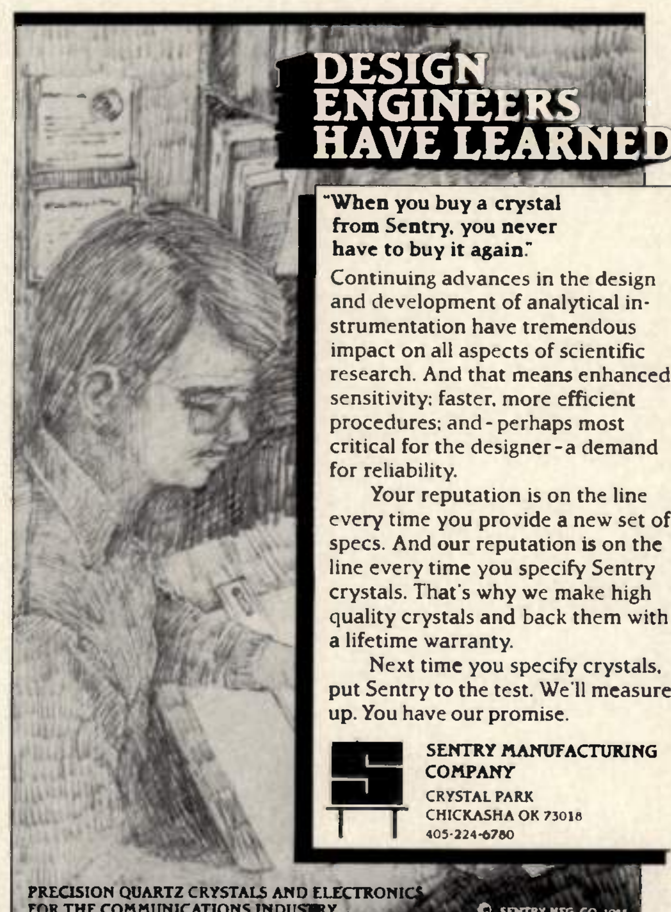
Included in a new 33-page Electrolytic Capacitor Catalog are complete electronic and mechanical specifications for high voltage and extended temperature range versions; miniature, subminiature, and super subminiature sizes; low leakage, low noise and bi-polar versions; low ESR and low impedance types. **Paccom Electronics, Redmond, Wash., INFO/CARD #142.**

Semiconductor Catalog

The comprehensive 400-page catalog from Alpha Industries contains technical data sheets and application notes for mixer and detection diodes, including Schottky barrier and point contact diodes; control diodes, including PIN switching, attenuator, and limiter diodes, and Silicon/GaAs tuning diodes; power generation devices, including Gunn diodes/modules, and Silicon GaAs parametric amplifier diodes; Silicon/GaAs multiplier and step recovery diodes. **Alpha Industries, Inc., Woburn, Mass., INFO/CARD #138.**

Triple Output Power Supply Comparison Chart

A free triple output power supply comparison chart covers Electronic Measurements, Lamdba, Kepco, Hewlett Packard, Dynascan and Systron Donner models. Parameters needed to select the proper triple output supply for your specific needs are detailed, including outputs for 1, 2, 3, max. output power outputs isolated, series operation, parallel operation, output modes, voltate current set, panel meter, simulation read volts and amps, regulation, ripple, isolation, output/case, frequency at output, size, bench area sq. ft., warranty and price. **Electronic Measuring, Inc., Neptune, New Jersey, INFO/CARD #137.**



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