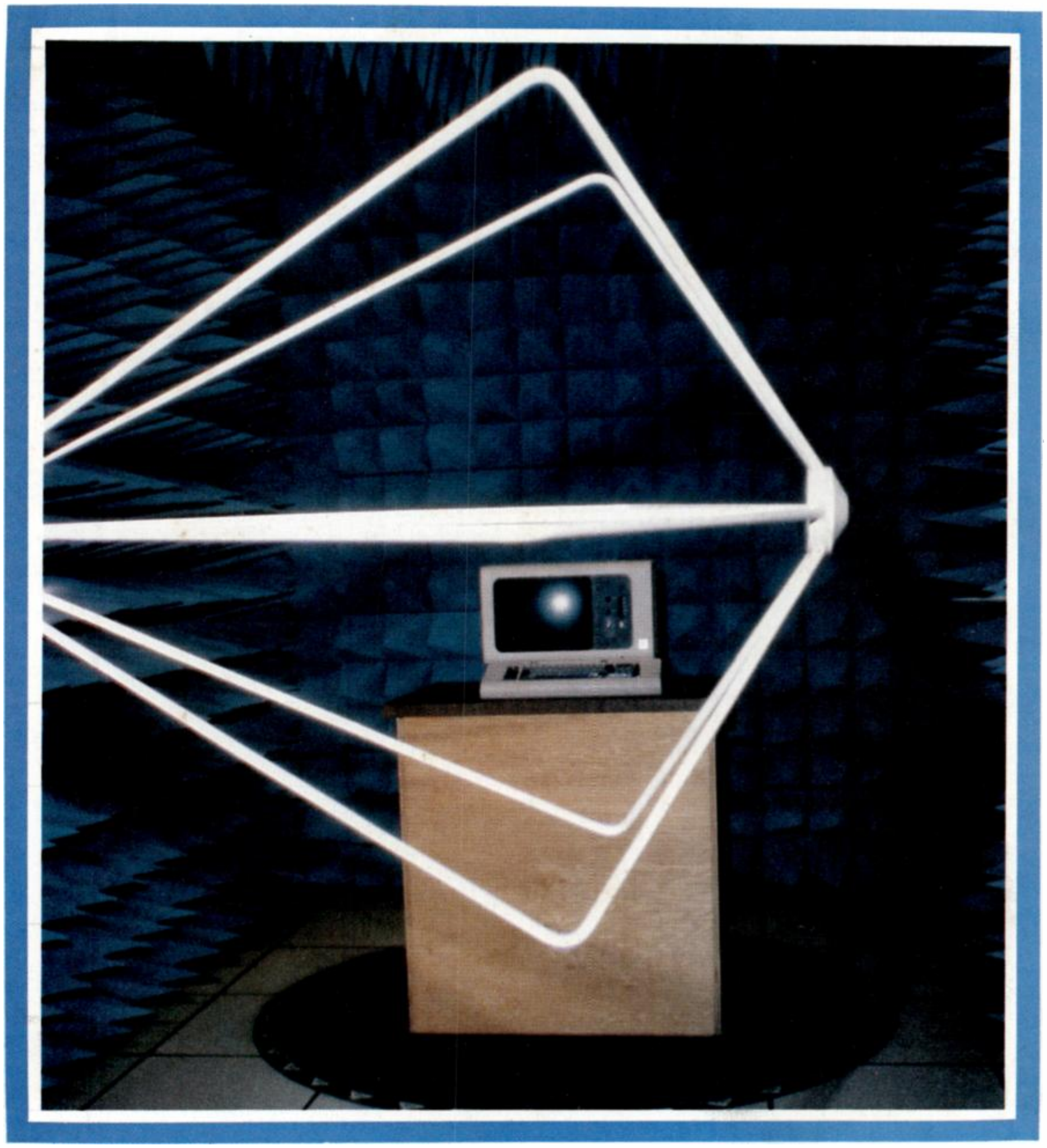
May/June 1981

A Cardiff Publication

rf design



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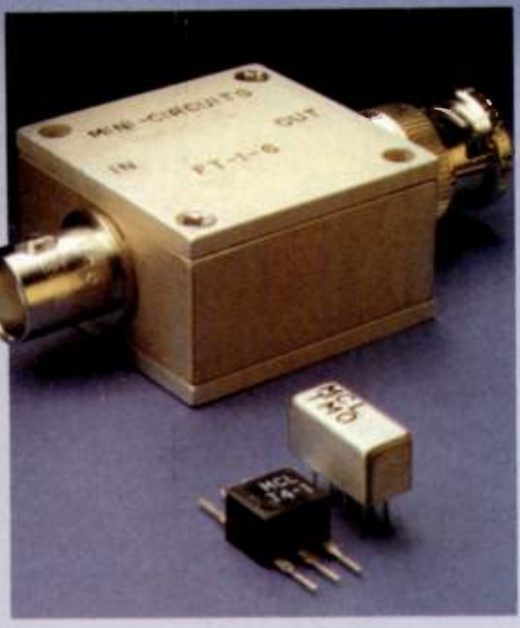

Select from the economical, microminiature T–series (plastic case) or TMO series (hermetically-sealed metal case) covering 10 KHz to 800 MHz. These models operate from 12.5 to 800 ohms with insertion loss typically less than 0.5 dB.

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DC ISOLATED PRIMARY & SECONDARY

Model No.	T1-1 TMO1-1	T1-1H	T1.5-1 TMO1.5-1	T2.5-6 TMO2.5-6	T4-6 TMO4-6	T9-1 TMO9-1	T9-1H	T16-1 TMO16-1	T16-1H
Imped. Ratio	1	1	1.5	2.5	4	9	9	16	16
Freq. (MHz)	.15-400	8-300	.1-300	.01-100	.02-200	.15-200	2-90	.3-120	7-85
T Model (10-49)	$2.95	$4.95	$3.95	$3.95	$3.95	$3.45	$5.45	$3.95	$5.95
TMO model (10-49)	$4.95		$6.75	$6.45	$6.45	$6.45		$6.45	

CENTER-TAPPED DC ISOLATED PRIMARY & SECONDARY

Model No.	T1-1T TMO1-1T	T2-1T TMO2-1T	T2.5-6T TMO2.5-6T	T3-1T TMO3-1T	T4-1 TMO4-1	T4-1H	T5-1T TMO5-1T	T13-1T TMO13-1T
Imped. Ratio	1	2	2.5	3	4	4	5	13
Freq. (MHz)	.05-200	.07-200	.01-100	.05-250	.2-350	8-350	.3-300	.3-120
T Model (10-49)	$3.95	$4.25	$4.25	$3.95	$2.95	$4.95	$4.25	$4.25
TMO model (10-49)	$6.45	$6.75	$6.75	$6.45	$4.95		$6.75	$6.75

UNBALANCED PRIMARY & SECONDARY

Model No.	T2-1 TMO2-1	T3-1 TMO3-1	T4-2 TMO4-2	T8-1 TMO8-1	T14-1 TMO14-1
Imped. Ratio	2	3	4	8	14
Freq. (MHz)	.025-600	.5-800	.2-600	.15-250	.2-150
T model (10-49)	$3.45	$4.25	$3.45	$3.45	$4.25
TMO Model (10-49)	$5.95	$6.95	$5.95	$5.95	$6.75

FT FTB

Model No.	FT1.5-1	FTB1-1	FTB1-6	FTB1-1-75
Imped. Ratio	1.5	1	1	1
Freq. (MHz)	.1-400	.2-500	.01-200	.5-500
(1-4)	$29.95	$29.95	$29.95	$29.95

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The 6070A and 6071A are equipped with an IEEE-488 bus interface as a standard feature for systems applications and all necessary front panel functions are remotely programmable. Complete status information can be supplied back to an instrument controller such as the Fluke 1720A. Both "teach" and "learn" modes are provided.

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RF 5/6-81

☐ Please send me complete 6070A specifications and applications literature.
☐ Send information on other IEEE products.
☐ Have a Sales Engineer contact me for a demonstration.

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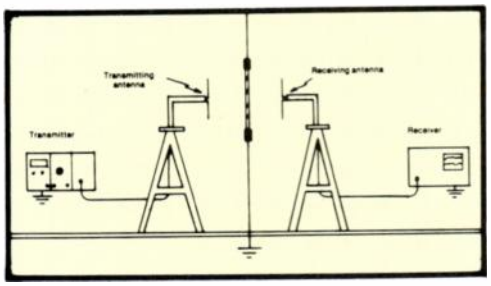
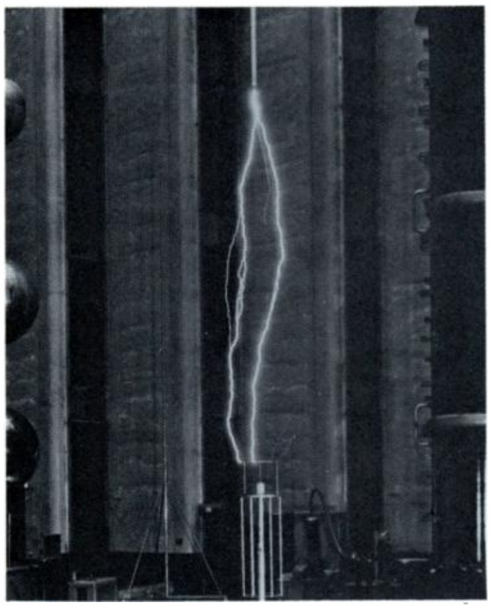
INFO/CARD 2

r.f. design

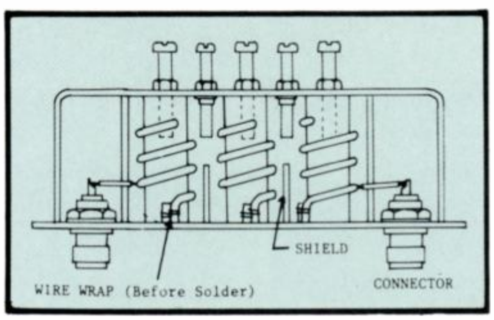
May/June 1981



EMC



Active Antennas



Helical Filters

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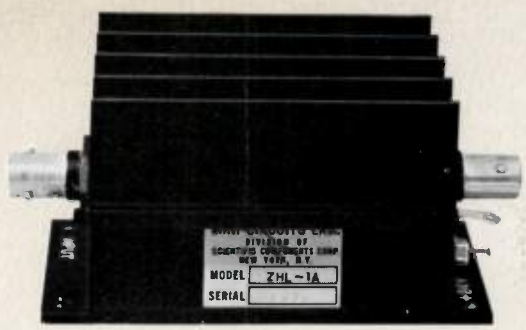
Using an ultra-linear Class A design, the ZHL is unconditionally stable and can be connected to any load impedance without amplifier damage or oscillation. The ZHL is housed in a rugged ⅛ inch thick aluminum case, with a self-contained hefty heat sink. BNC connectors are supplied; however, SMA, TNC and Type N connectors are also available. Of course, our one-year guarantee applies to each amplifier.

So from the table below, select the ZHL model for your particular application now ... we'll ship within one week!



ZHL-1A

MODEL NO.	FREQ. MHz	GAIN dB	GAIN FLATNESS dB	MAX. POWER OUTPUT dBm 1-dB COMPRESSION	NOISE FIGURE dB	INTERCEPT POINT 3rd ORDER dBm	DC POWER VOLTAGE	DC POWER CURRENT	PRICE $ EA.	PRICE QTY.
ZHL-32A	0.05-130	25 Min.	±1.0 Max.	+29 Min.	10 Typ.	+38 Typ.	+24V	0.6A	199.00	(1-9)
ZHL-3A	0.4-150	24 Min.	±1.0 Max.	+29.5 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199.00	(1-9)
ZHL-1A	2-500	16 Min.	±1.0 Max.	+28 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199.00	(1-9)
ZHL-2	10-1000	15 Min.	± 1.0Max.	+29 Min.	18 Typ.	+38 Typ.	+24V	0.6A	349.00	(1-9)
ZHL-2-8	10-1000	27 Min.	±1.0 Max.	+29 Min.	10 Typ.	+38 Typ.	+24V	0.65A	449.00	(1-9)
ZHL-2-12	10-1200	24 Min.	±1.0 Max.	+29 Min.*	10 Typ.	+38 Typ.	+24V	0.75A	524.00	(1-9)

Total safe input power +20 dBm, operating temperature 0° C to +60° C, storage temperature −55° C to +100° C, 50 ohm impedance, input and output VSWR 2.1 max.
*+28.5 dBm from 1000-1200 MHz
For detailed specs and curves, refer to 1980/81 MicroWaves Product Data Directory, Gold Book, or EEM.

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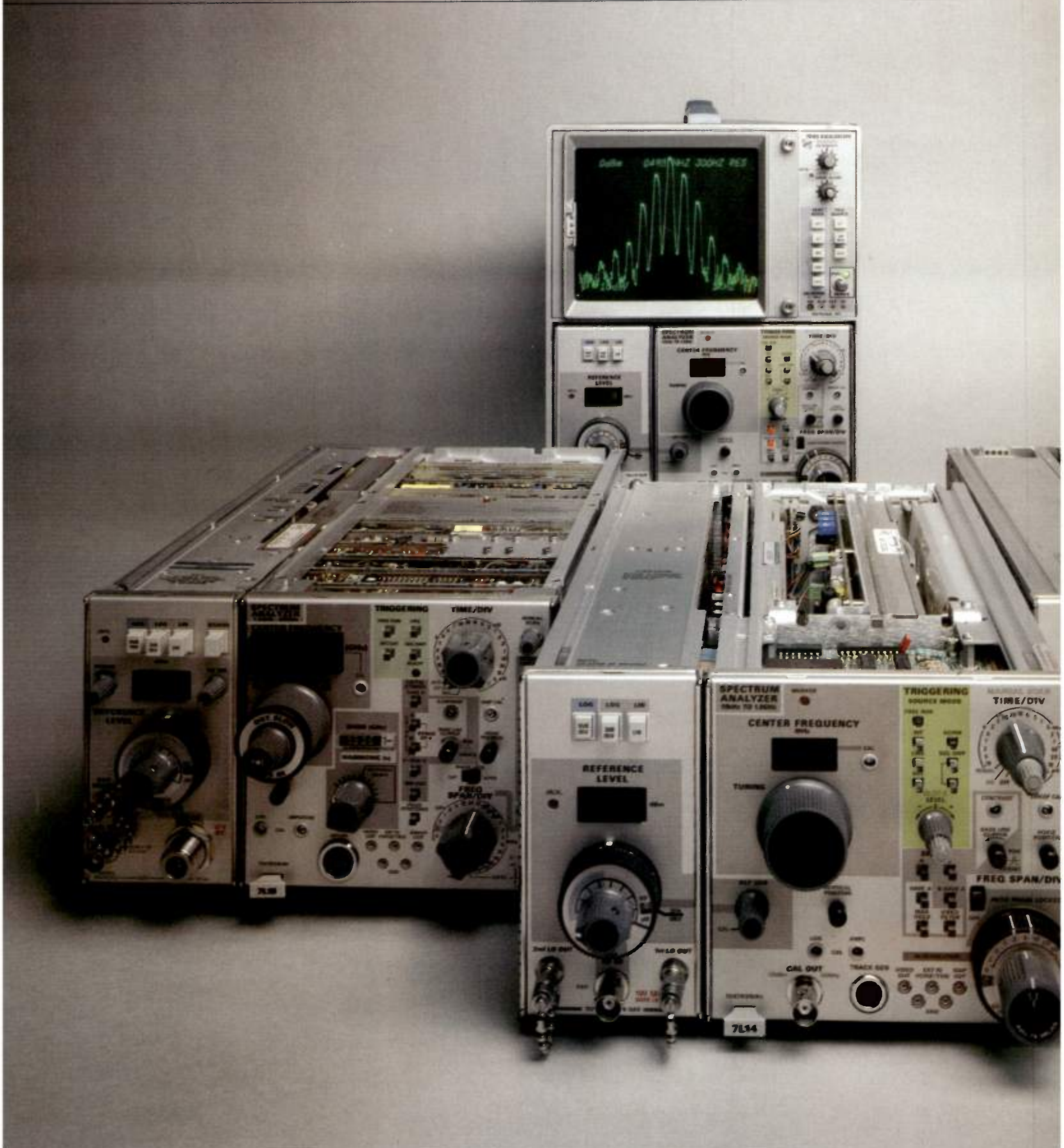
INFO/CARD 3

TEK 7000 SERIES SPECTRUM ANALYZERS



Now you can plug in an all digital-storage family of spectrum analyzers.

Three models provide laboratory performance with flexibility and versatility.

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Family characteristics that make spectrum analysis easier.

All three instruments feature digital storage for flicker-free displays that are easy to interpret. This provides averaging and peak detection; accurate waveform comparisons; stores for long periods to measure amplitude changes and frequency drift.

The 7000 Series plug-in family displays alphanumeric readout for referencing and easy documentation. And each spectrum analyzer is protected from up to one watt input levels to save expensive front end repairs caused by inadvertent overloading.

With a 7000 Series mainframe on your bench, you select the spectrum analyzers that fit your requirements. And they interchange quickly with 30 other Tektronix test and measurement plug-ins. Use the powerful mainframe for logic analyzer, oscilloscope and other measurements.

New 10 KHz to 1.8 GHz 7L14 completes the 20 Hz to 60 GHz digital storage plug-in team.

Here it is. The new 7L14 for digitally-stored close-in, high resolution measurements from 10 KHz to 1.8 GHz. With 10 Hz residual FM, the 7L14 provides stability and jitter-free 30 Hz resolution displays. Its digital storage can be used to eliminate system errors and provide flat swept RF measurement capability. Digital averaging provides noise reduction which gives 70 dB spurious-free dynamic range. You can check broadband RF networks, filter networks, amplifiers, cables. Measure EMI/RFI and FM, navigation, two-way and other communications systems.

At the top of the spectrum you get top performance from the 7L18 Spectrum Analyzer. It provides full amplitude calibration in the 1.5 GHz to 60 GHz range and has 30 Hz resolution to 12 GHz. Displays are sharp, stable and flicker-free. Digital storage and digital signal processing make complex measurements easy with microprocessor aided controls. An automatic preselector insures spurious-free operation, giving easily interpreted displays.

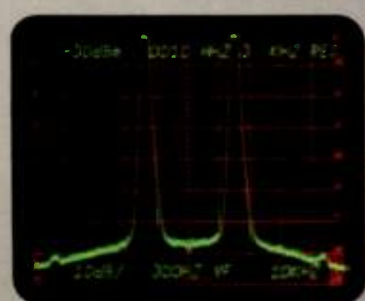
For baseband measurements choose the 7L5 for its precision and convenience in the 20 Hz to 5 MHz range with 10 Hz resolution. For a high performance analyzer, it's unusually easy to operate.

This family works together to make an outstanding value.

You get this laboratory performance and measurement flexibility at prices that point up the value of the Tektronix plug-in concept.

Call your nearest Tektronix Field Office (listed in major city directories) for complete details on the 7000 Series lab performance spectrum analyzers. Or call 800-547-1512 for descriptive literature.

Tektronix, Inc., P.O. Box 1700, Beaverton, OR 97075. In Europe: European Marketing Centre, Postbox 827, 1180 AV Amstelveen, The Netherlands.



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OH-OHH

Hey what gives? A quick scan of the issue and what do you see? "Computing devices. . . ." I thought this magazine was going to remain RF and not get into the bits and bytes like most of the other "analog" books have? The answer is we're still RF — they've joined *us*.

The FCC recently (via Docket 20780) has taken a closer look at the radiated and conducted emissions of some computing devices and decided that a ruling is called for. It appears that timing signals or pulses greater than 10 kHz running along unshielded paths have a tendency to radiate energy in the RF spectrum. I'm sure any RF designer worth his salt could have foreseen this as a problem well before the recent FCC rulings were considered.

As a matter of fact several RF engineers have helped define the problem areas and effect some cures in this issue. Leonard Levin of R&B Enterprises, Iver Sonderby of Stanford Applied Engineering, and Jim Coniglio of Acheson Colloids provide several EMI/RFI/EMC points-of-view in the following pages.

Once again I'm pleased to present subject matter in an area that has received relatively little attention in the open literature to date. Dr. Rohde presents in this issue an article on "Active Antennas" from a component through system point-of-view.

Also included in this issue and certainly no less important are in-depth articles on ceramic/porcelain multilayer capacitors, helical filters and oscillators.

Rounding out this issue is a book review section on some of the indispensible books in our field. If you don't have them, they are worth obtaining for your working library.

Speaking of shows (we were?), Electro '81 in New York City was an experience. We had a booth on the third floor at the Coliseum. To those who stopped by to say hello — thank you for your kind words. Hopefully we'll see many more of you at the MTT and Wescon '81 later this year. Space and interest permitting, look in this and subsequent issues of *r.f. design* for future electronic trade show announcements.

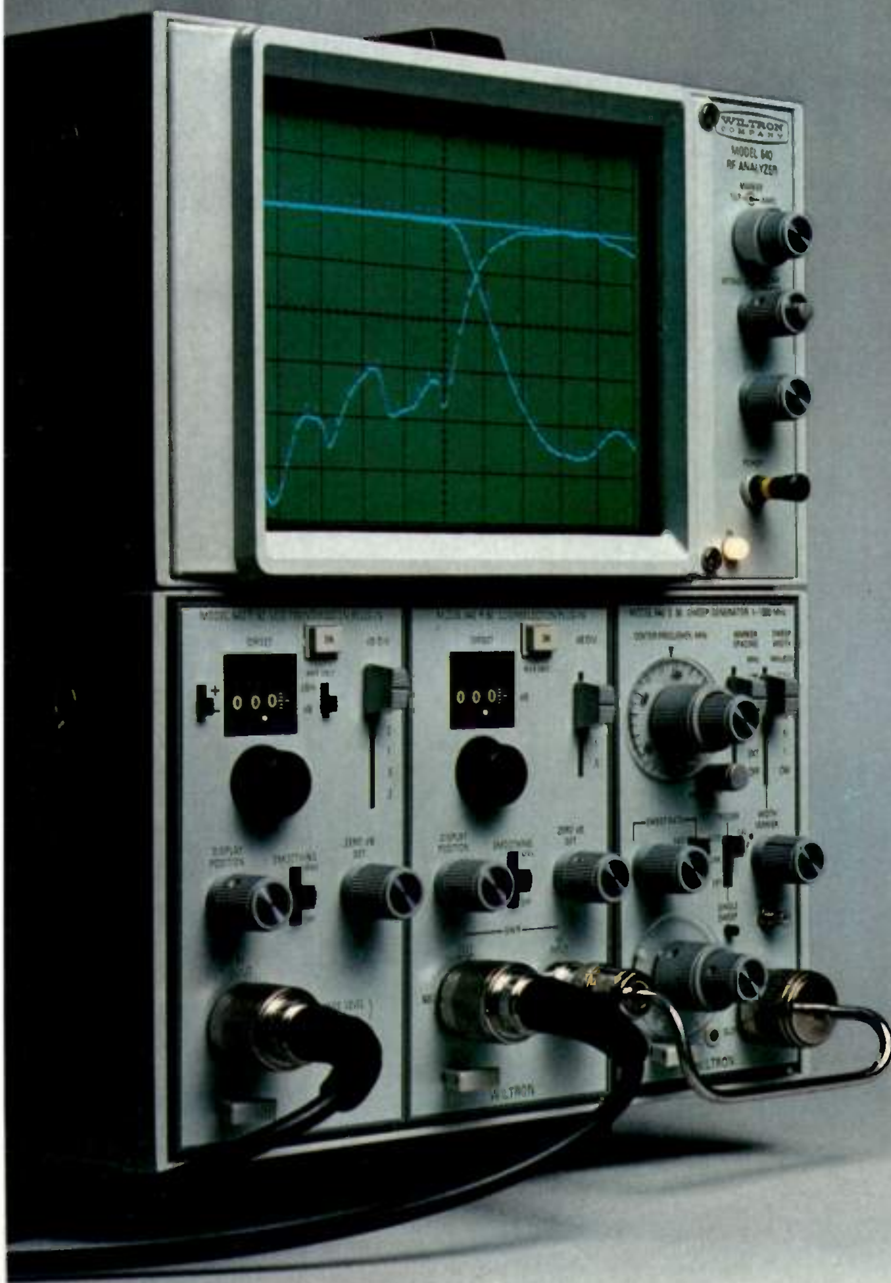
P.S. If you know of any shows that quite a few of us might be interested in, drop me a line or call.

P.P.S. Hey — get out there and vote — oops! wrong topic. Fill out the qualification/subscription card and send it in pronto! Please. *Everyone* — qualified must be counted.

Rich Rosen

Measure transmission and reflection with stark simplicity, 1-1500 MHz.

It's a Wiltron.



Here in one compact instrument, Wiltron gives you a complete RF Analyzer for swept measurements on 50Ω or 75Ω devices over the 1 to 1500 MHz range.

Use the Wiltron 640 to make transmission gain/loss, reflection (return loss/SWR), absolute power and absolute frequency measurements. You'll find the 640 is one of the easiest instruments you've ever used. Simply connect the test device. You won't need an armful of couplers, amplifiers, cables or other equipment. All the circuitry – sweeper, directional signal separator, calibrated amplifiers, detectors and display system – is inside the case.

No more muddled measurements.

Wiltron's 640 offers features you won't find in far bigger, more expensive instruments.

It gives you the most versatile marker system available, stable, crystal accurate, on a dual-trace display with a ±90 dB calibrated offset.

You can precisely sweep over the entire 1500 MHz range or over just 1 MHz. Dynamic measuring range is 70 dB (+15 to −55 dBm). Measure return loss to below 30 dB (1.06 SWR).

Low cost plug-in flexibility.

Five plug-ins are offered: a swept signal source, log reflection unit, log transmission unit, linear amplifier unit and log transmission/reflection unit for use with external detector and SWR Autotester.

Ask for Technical Review #7.

For complete details ask for a copy of Technical Review #7. For an early demonstration, call Walt Baxter, Wiltron, 825 East Middlefield Road, Mountain View, CA 94043. Phone (415) 969-6500.

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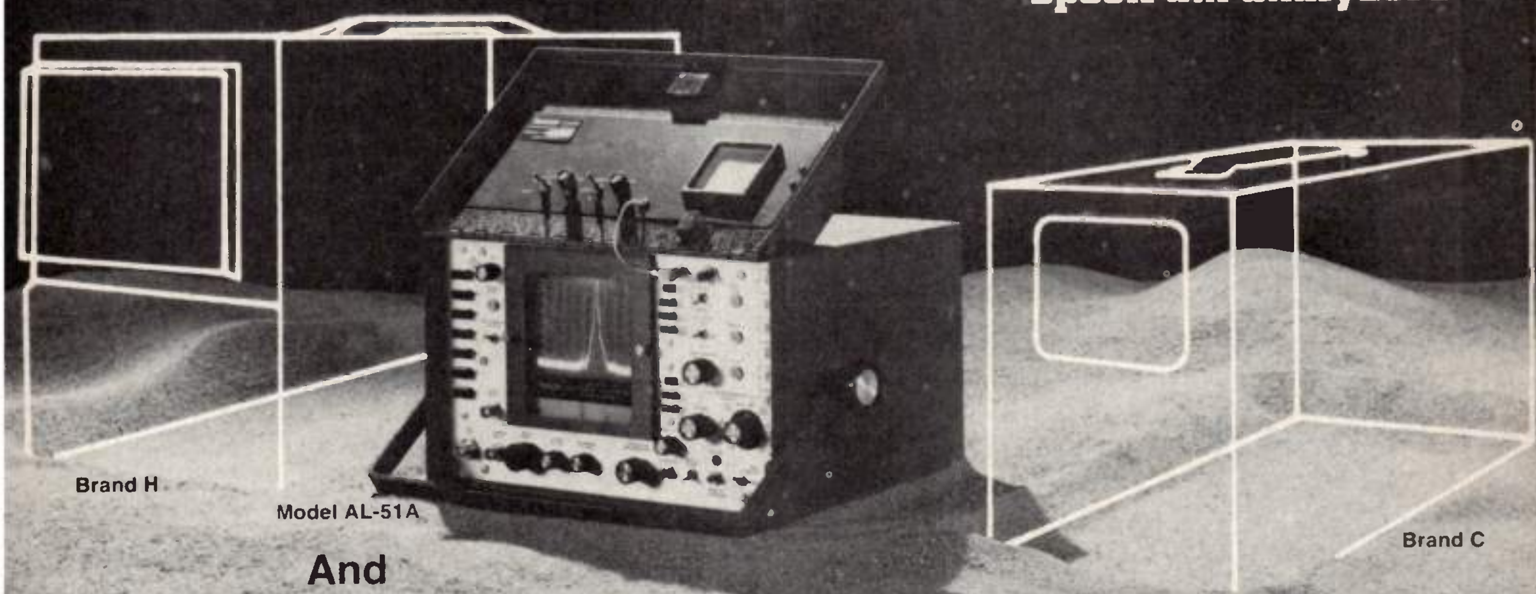
INFO/CARD 5

the rugged truth

Texscan is the world leader in portable spectrum analyzers



And here's proof . . .

SPECIFICATION	BRAND H*	TEXSCAN'S AL51A	BRAND C*
PRICE	$7975.00	$5295.00	$6255.00
FREQUENCY RANGE DISPERSION FREQUENCY ACCURACY	0.1—1500 MHz 50KHz—1000MHz 2% of dispersion ± 5MHz	0.4—1000MHz 20KHz—1000MHz ± 0.01%	1—1000MHz 100KHz—100MHz ±5MHz
AMPLITUDE DYNAMIC RANGE AVERAGE NOISE LEVEL ACCURACY (total worst case)	70dB −107dBm (10KHz resolution) ±3.5dB	60dB −118dBm (10KHz resolution) ±4dB	70dB NOT SPECIFIED ±3dB
RESOLUTION (min)	1KHz	500Hz	2KHz
STABILITY SHORT TERM P/P LONG TERM	 1KHz NOT SPECIFIED	 500Hz 25KHz/10 min	 NOT SPECIFIED 50KHz/5 min
NOISE SIDEBANDS	−65dB 50KHz away	−70dB 50KHz away	−70dB 50 KHz away
OPERATING POWER	115/230vac	115/230vac 12 vdc	115/230 vac
SIZE CUBIC INCHES	2059	1092	1503
WEIGHT LBS	40lbs	27lbs (incl battery)	30lbs
FEATURES	BRAND H*	TEXSCAN'S AL51A	BRAND C*
INTERNAL BATTERY	NOT OFFERED	STANDARD	NOT OFFERED
EXTERNAL 12v ac oper	NOT OFFERED	STANDARD	OPTIONAL
PHASELOCK	NOT OFFERED	STANDARD	NOT OFFERED
AUDIO	NOT OFFERED	OPTIONAL	STANDARD
FREQUENCY MARKERS	NOT OFFERED	STANDARD	NOT OFFERED
DIGITAL STORAGE	OPTIONAL	OPTIONAL	NOT OFFERED
PRESET FREQUENCY BANDS	NOT OFFERED	STANDARD	NOT OFFERED
TWO LOG RANGES	STANDARD	STANDARD	NOT OFFERED
RUGGED CARRYING CASE WITH FRONT PANEL COVER	OPTIONAL	STANDARD	OPTIONAL

*Information obtained from manufacturer's published data

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INFO/CARD 6

How to Cope With the New FCC Rules Regarding Computing Devices

Leonard Levin
R&B Enterprises
Plymouth Meeting, PA

The hand-held microprocessor-controlled calculator made its debut in the mid-sixties and, by the early seventies, began to come into common usage. By the mid to late seventies, the personal computer was in general use and other microprocessor-controlled devices, such as home appliances and TV games, as well as automotive control systems, also were becoming commonplace. The vast proliferation of these products has been an electronics revolution; but, the side effect has been a growth in environmental electronic noise. The microprocessor is a tiny component capable of performing many complex operations using "digital" techniques; but, resultant high frequency signals can be transmitted from the device, and the resulting uncontrolled electronic noise has the potential to disturb communications and other unprotected microprocessor-controlled devices.

Docket 20780

Government and military organizations long have recognized the problem of uncontrolled electronic noise, and recently the FCC issued a strong set of regulations, Docket 20780, and a timetable for compliance. This action, which is comparable to the automotive emission restrictions of a decade ago, has had a tremendous impact on manufacturers of microprocessor-controlled devices, as they now must reduce or eliminate the electromagnetic interference generated by their products, whereas, in the past, electromagnetic radiation may not even have been considered.

The basic intent of these new regulations, which are actually restrictions, is to provide a reasonable (electronic) pollution-free environment. This is accomplished by controlling the level of radiated and line-conducted emissions emanating from the devices.

What is A Computing Device?

A computing device is defined as any electronic device or system that generates and uses timing signals or pulses at a rate in excess of 10,000 pulses (cycles) per second and uses digital techniques. Additionally, the FCC regulations require that some other terms be understood. These are:

Class A Computing Device: Marketed for use in a commercial, industrial, or business environment.

Class B Computing Device: Marketed for use in a residential environment, notwithstanding use in a commercial, industrial, or business environment.

Verification: A procedure where the manufacturer tests the equipment and takes the necessary steps to insure that the equipment complies with the appropriate standards. Submittal of a sample unit or test data to the FCC is not required unless specifically requested; but, test data should be retained on file as proof of compliance.

Certification: An equipment authorization issued by the FCC based on test data submitted by the applicant. Certification authorizes the manufacture and sale of the product.

As a rule, most Class A devices are located within commercial premises and are subject to a good level of maintenance. Class B devices are commonly located in residential premises, close to radio, TV and similar equipment. Therefore, in general, the more strict certification procedure is required for Class B devices, while the verification procedure is applicable for Class A devices. However, the FCC[1] can advise you whether certification or verification applies to a specific product.

Labeling Requirements

Labeling requirements are also specified. Briefly, they require that all equipment placed in production prior to Jan. 1, 1981 must be labeled as of Jan. 1, 1981 to state whether the equipment has been tested and whether it may cause interference. Any new equipment placed into manufacture to be released in 1981, and thereafter, must meet and prove compliance by Oct. 1, 1981. There are several other dates that are specified, and these may be determined best through the FCC.

Results Of *Non*-compliance

These new restrictions on electromagnetic interference are backed by the full power of the FCC and non-compliance can result in fines and imprisonment. The FCC also can order the cessation of manufacture and sales of non-compliant equipment. Consideration of these restrictions has brought the science of electromagnetic compatibility (EMC) new publicity, and has elevated the EMC engineer into new prominence.

EMC is the ability of an electronic device to operate within its intended environment, and it usually deals with the reduction of RFI/EMI emissions from an electronic device and also reducing the susceptibility to RFI/EMI. This is best accomplished during the original design of equipment; but, where necessary, EMC techniques can be used effectively on existing equipment.

EMC Techniques

These techniques are varied, and are well known to the experienced EMC engineer, as are the methods for testing for their effectiveness. The desired overall end result is that radiated and conducted interference be at or below the specified required levels. The techniques consist of two generalized categories: *filtering*, to reduce or to eliminate line conducted interference; and, *shielding*, to protect against electromagnetic radiation. There are various methods to accomplish filtering and shielding, and their results can be two-fold. One is to reduce the output of undesirable electromagnetic by-products from equipment; the other is to protect electronic equipment from outside electromagnetic interference, thus enabling the equipment to operate in an otherwise electromagnetically unfavorable environment.

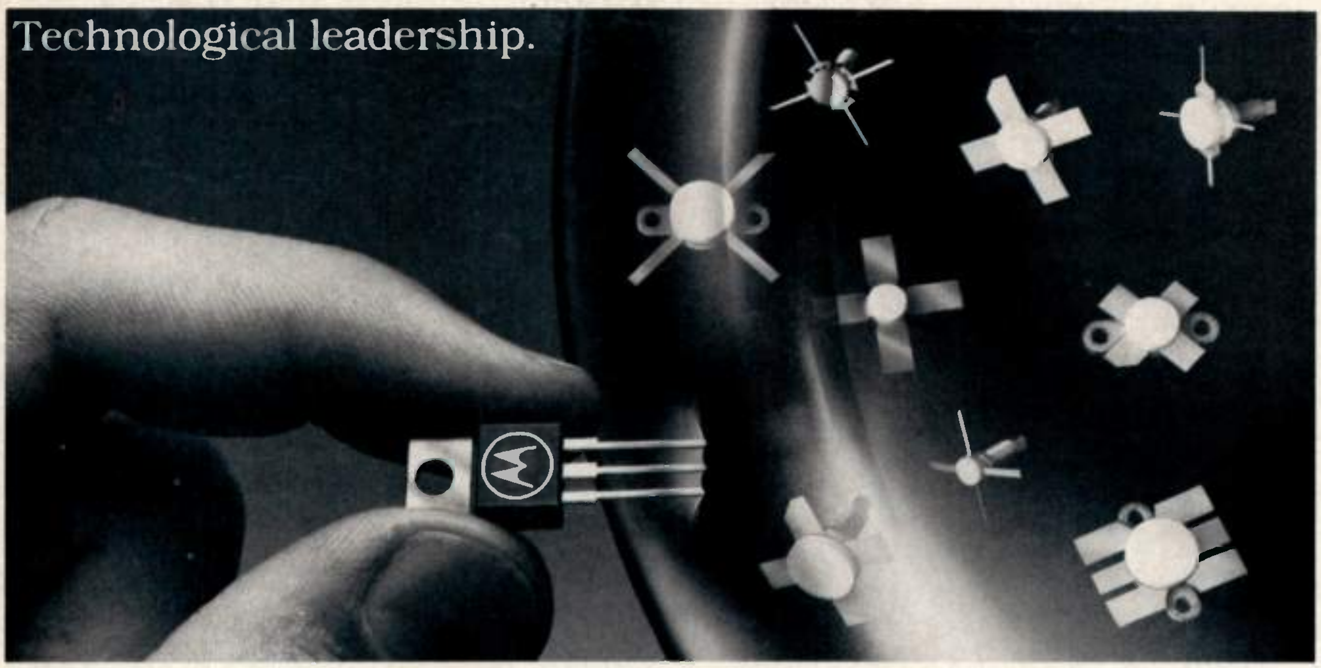
A non-electrical solution to EMI, where conventional shielding and filtering techniques are not practical or effective, is the use of fiber optics; but, this is a complete subject in itself.

Where To Go For Help?

Seminars are given that specifically address the new FCC Dockets and their consequences on design and testing. Others cover military standards, European VDE specifications and product safety; or, just general categories of EMC. Additionally, there are consulting and testing organizations which specialize in EMC and provide services ranging from consultations with an EMC design engineer to complete FCC testing, including all reports and paperwork required by the FCC. □

Reference

[1]Federal Communications Commission, RF Devices Branch (Authorization and Standards Division), Washington, DC 20554; (202) 653-8121 and 1828.



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Concentration on SOE package performance for more gain, bandwidth and P_{out} at higher and higher f_T has brought device costs right up with it. And, while the widely-accepted, SOE package has excellent RF performance, it contains substantial gold, beryllium oxide and kovar — rare metals and materials that have further inflated the cost bubble to the breaking point.

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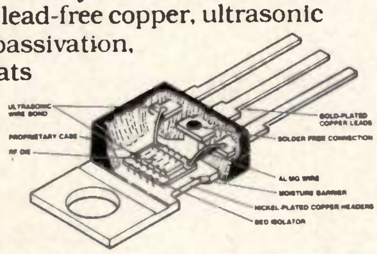

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We offer 14 devices, most common-emitter configuration, with flanges and center lead emitter-connected for optimum mounting and heatsinking. More are in the wings.

DEVICE TYPE	P_{out} WATTS	G_{PE} dB MIN	V_{CC} VOLTS	PRICE 100-499
1.5-30 MHz HF/SSB/AM				
MRF476*	3	15	12.5	1.75
MRF475*	12	10	13.6	2.50
MRF485*	15	10	28.0	3.00
MRF477	40	15	12.5	10.00
MRF486	40	15	28.0	12.00
25-50 MHz: VHF LB				
MRF479	15	12	12.5	8.00
MRF497	40	10	12.5	10.00

DEVICE TYPE	P_{out} WATTS	G_{PE} dB MIN	V_{CC} VOLTS	PRICE 100-499
100-150 MHz: VHF AM				
MRF340	8	13	27.0	4.50
MRF342	24	11	27.0	8.00
MRF344	60	8	27.0	11.00
136-175 MHz: VHF HB				
MRF260	5	10	12.5	4.00
MRF261	10	5.2	12.5	4.50
MRF262	15	6.3	12.5	8.00
MRF264	30	5.2	12.5	9.00

*Collector Tab — All Others Are Emitter Tab

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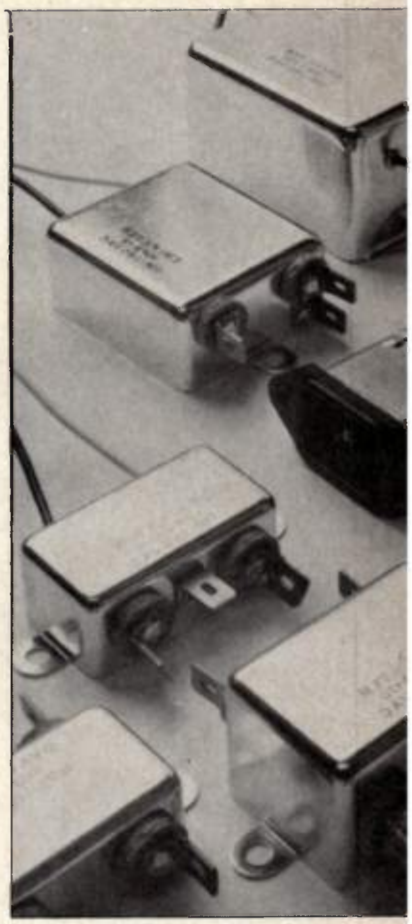
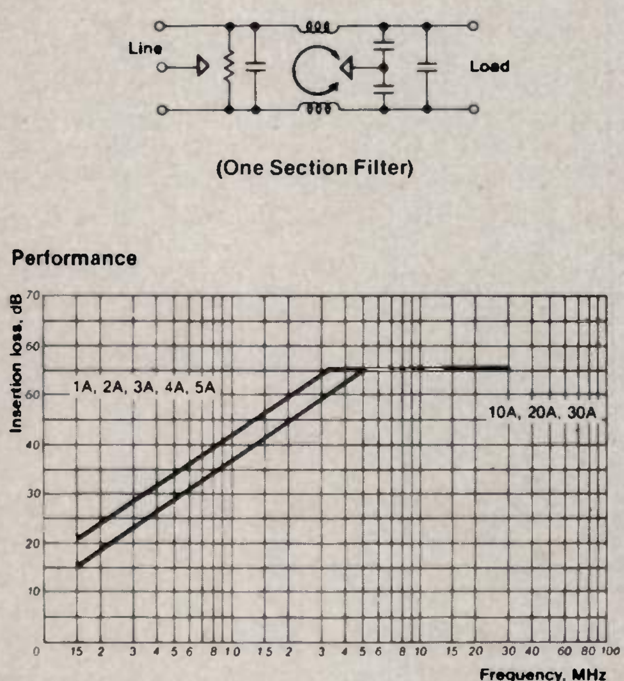
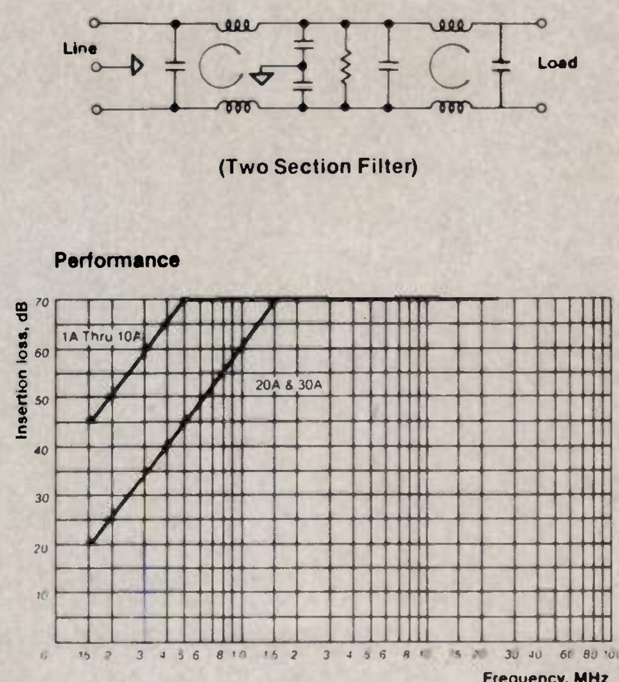


Typical EMI line filters.



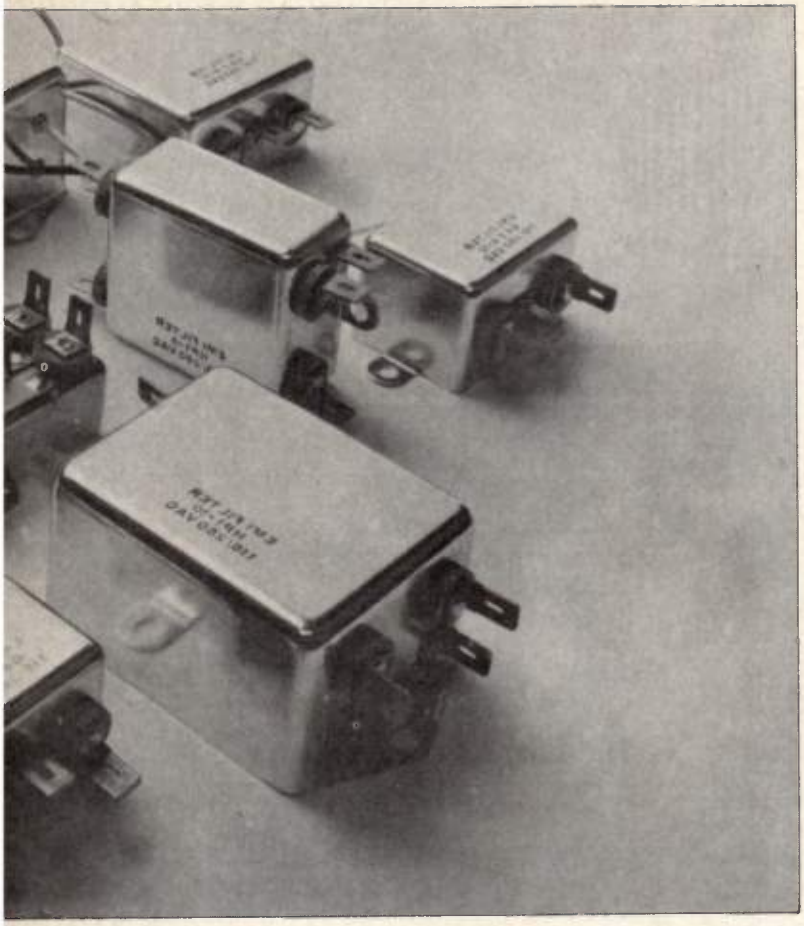
Insertion Loss in 50 Ohm System typ.
per MIL-STD-220A

Figure 1.



Insertion Loss in 50 OHM System typ.
per MIL-STD-220A

Figure 2.



EMI/RFI Filtering Of Computing Devices

Iver Sonderby
Stanford Applied Engineering
Santa Clara, CA

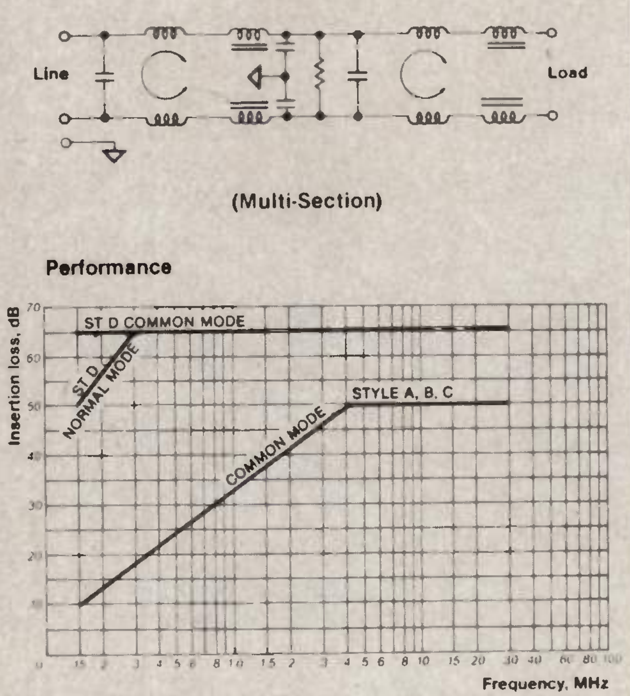
On Oct. 11, last year, the Federal Communications Commission adopted regulations to restrict and reduce the interference potential of electronic computing equipment. The rules and regulations are outlined in Docket No. 20780, which define a computing device as "any electronic device or system that uses digital techniques, or more precisely, an electronic product that intentionally generates and uses radio frequency in excess of 10,000 cycles or pulse per second".

The definition of "computing device" is intentionally broad and will encompass any computation equipment used for control, transformations, recording, filing, sorting, storage and retrieval, data terminals, word processors and similar products. Other devices subject to the new regulations are RF power supplies, electronic games, carrier current systems, campus radio stations, calculators, and tape recorders.

The New Regulations

What effect will these new regulations have on present EMI/RFI filtering techniques? Before the October 11 adoptive of the FCC regulations, manufacturers were utilizing EMI/RFI filters to protect their equipment from "conducted interference" emanating from the power line that could cause malfunctions in microprocessors or other noise-sensitive circuitry. Filters of the "single network" type (Figure 1) were popular and widely used because of their low cost, small size, and their ability to suppress adequate noise and interference from the power line. Little or no concern was given to the equipment itself and to the internal interference generated, which is of sufficient intensity to pass back through the filter at a level that is now unacceptable to the FCC. Herein lies the problem. Many of these manufacturers must now upgrade their filtering to the more expensive, and larger, "two network" filters (Figure 2). And if a switching power supply is utilized, they are required to use a multi-section filter (Figure 3) to adequately suppress the high-intensity noise generated by such a device. Altering equipment to facilitate a larger filter is relatively easy to accomplish, but the re-design of enclosures and shielding necessary to meet conducted radiation limits is a far more serious problem and is far more costly.

Interestingly, the FCC followed the German VDE



Insertion Loss in 50 Ohm System Typ. per MIL-STD-220A

Figure 3.

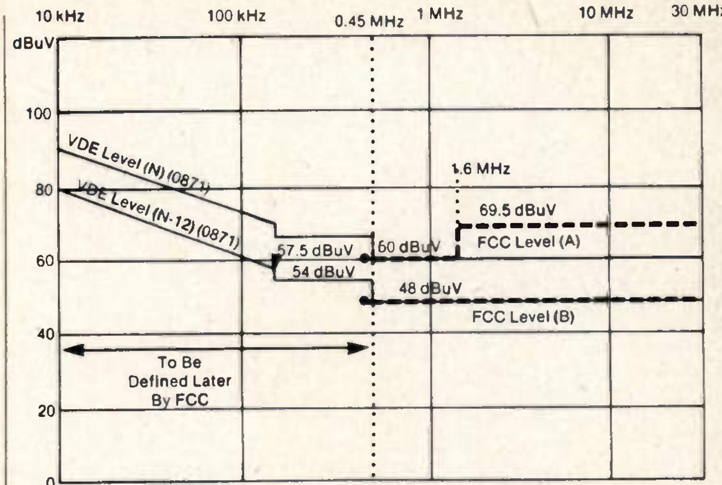


Figure 4. Comparison-Conducted Noise: FCC 20780 VS VDE 0871.

specification (0871) very closely. Figure 4 shows the relationship between the two and the differences between allowable noise levels at various frequencies. The FCC "Level A" curve applies to computing devices used in an "industrial environment"; the "Level B" curve applies to computing devices, electronic games, or any device intended to be connected to a TV receiver or TV interface device. Excluding the use of a switching power supply in the system, the "Level A" requirement can most readily be met through use of a "two-section filter" (Figure 2). If a switching power supply *is* used, a multi-section filter is recommended (Figure 3) for maximum noise attenuation.

"Level B" is by far the most stringent, and in many instances a two-section filter will not suffice. There are a number of commercial "high-performance" filters available that may supply the suppression required. However, if a switching power supply is incorporated within the equipment, it may require a custom design to bring the interference level below the 250 microvolts limit from .45 MHz to 30 MHz.

Recent Conducted Emission Tests

Numerous conducted emission tests on various types of computing devices have been performed* since the new FCC ruling came into effect on Oct. 1. Findings show that equipments which use filters of the "single network" type (Figure 1) are marginal in suppressing emission to meet the "Level A" requirement in some cases. Data terminals that had an internal "clock rate" of 1 MHz to 3 MHz have a tendency to exceed the maximum levels allowed at those frequencies.

Installation of a "two-section" filter (Figure 2) brings the level down significantly, to where conformance is acceptable. To meet the "Level B" requirement in one instance, it was necessary to design a custom, multi-section filter to adequately suppress conducted emission from a home computer that incorporated a switching power supply.

Needless to say, these new FCC regulations are causing a great deal of concern in the computer industry. On the brighter side, there are a number of EMI/RFI filter manufacturers, who have the expertise and readily available products to solve many of the problems that have arisen since the new FCC ruling was adopted. □

*By Stanford Applied Engineering.

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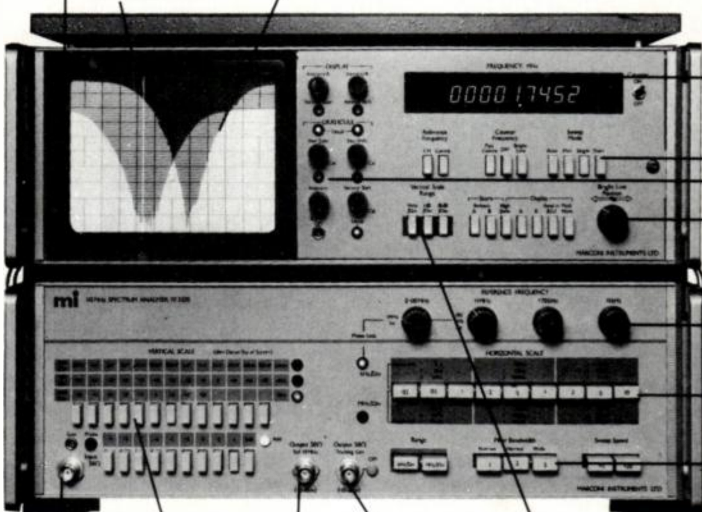

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INFO/CARD 9

EMC: A Problem

J.J. Coniglio
Acheson Colloids Co.
Port Huron, Michigan

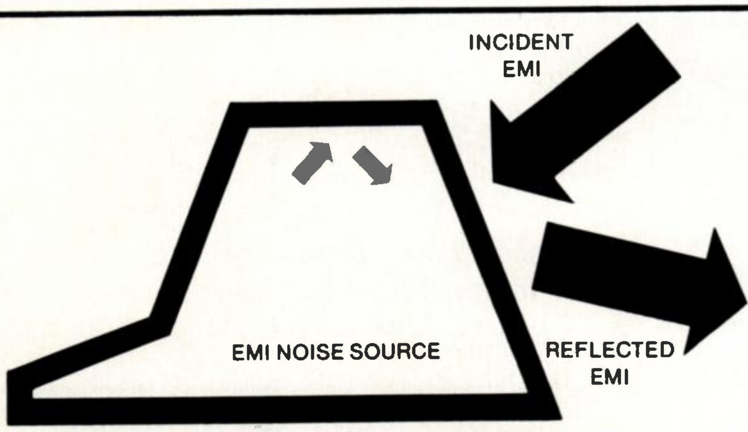

Figure 1. Electrical device with metal housing.

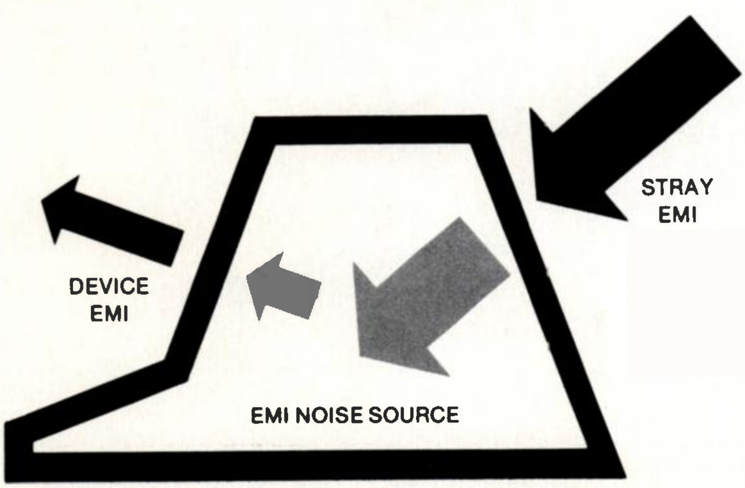

Figure 2. Electrical device with unshielded plastic housing.

The ability of electrical devices to function normally without being interfered with, or without interfering with other electrical devices, is what is thought of as Electromagnetic Compatibility. EMC regulations usually emphasize *containment* of electromagnetic interference (EMI) to specific levels across designated frequency ranges. Every electronic system has some level of electromagnetic radiation associated with it. If this is sufficiently strong enough to cause other equipment to malfunction, the radiating device is considered a noise source and is usually subjected to shielding regulations. This is especially true when the EMI occurs within the normal frequencies of communication as with video games and personal computers.

Shielding Effectiveness

Provided that the electronic system itself has been properly designed, the metal housing containing it shields by reflecting interfering signals away from the enclosure, as well as containing any EMI which may be radiating from the system itself. See Figure 1.

An unshielded plastic housing enclosing the same device would allow the radiating EMI to exit the housing as well as allow the entry of any stray EMI which may be present. This is shown in Figure 2.

The electrical energy present in the

Whose Time Has Come

environment, or radiating from an electrical device can be measured with various detectors, e.g. receivers and expressed in terms of standards of electrical measurement. There are volts/meter and watts/meter[1] for the electric field. Differences in levels of electrical signals or EMI are expressed in decibels as shown below as shielding effectiveness (SE):

$$SE = 20 \, Log \, \frac{V\,incident}{V\,transmitted} \, dB$$

or

$$SE = 10 \, Log \, \frac{P\,incident}{P\,transmitted} \, dB$$

The above relationships take into consideration both the reflected and absorbed components of the incident wave of EMI and are then the total shielding effectiveness (SE) for any material under examination at a *particular frequency.* Electrical energy impinging upon a conductive surface is reflected and absorbed by the surface as a function of the frequency and surface impedance. Table 1 shows the amount of electric field energy transmitted through a test panel expressed in dB Reduction, % Signal Reduction, and % Signal Transmitted.

Figure 3 shows the idea of reflection and absorption of EMI on plastic which has been shielded by some means.

The actual amount of additional shielding a system will require, depends on how much attenuation is needed to meet a specific emission standard, how susceptible the equipment is to ambient (existing) EMI, and how sensitive the manufacturer is to his costs in field service repair or his lost business due to EMC related performance problems in the field.

This is an edited version of part of the proceedings of the S.P.I. 7th structural foam conference of May 1979, courtesy of Technomic Publishing Co., Westport, Conn.

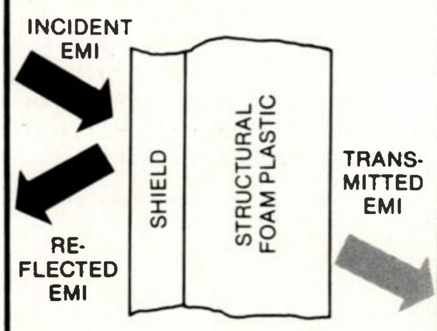

Figure 3. Shielded plastic.

Table 1.

dB Reduction	% Signal Reduction	% Signal Transmitted
0	0	100
3	50	50
6	75	25
9	87.5	12.5
12	93.75	6.25
15	96.75	3.125
18	98.44	1.56
21	99.22	0.78
24	99.61	0.39
27	99.81	0.19
30	99.91	0.09

Figure 4.

Background — ESD

Electrostatic discharge (ESD) affects the plastic industry in much the same way that EMI does. Plastics are generally excellent insulators and therefore do not allow charges to bleed off to ground in a controlled manner. Instead, discharges to the plastic case in many instances can cause equipment malfunction or failure. The ESD problem is becoming much more acute due to the development of very high-speed microprocessors. More and more functions are being packed onto a single chip to reduce the size of the equipment and increase machine speed.

A basic computer functions by communicating on data busses with very precise timing. When an improper signal is superimposed on the timing or other data busses within the computer, errors result. If the induced voltage generated by direct discharge is of sufficient magnitude, device malfunction (error) may occur as is shown in Figure 4. The H's are shown in normal and abnormal condition as a function of direct discharge testing at various voltage levels made directly to the cabinet of a dot matrix printer. As little as 100 volts can degrade several types of semiconductor devices commonly used in microprocessors. These devices work at very low signal levels. Therefore, very low voltage differences represent the 0 to 1 state, and they are very easily damaged even in their own manufacturing processes unless precautions are taken to protect them from ESD.

Testing Methods — EMI

EMI testing can be done on the materials under consideration for use as shielding and must be performed on the individual electronic equipment undergoing design. The shielding effectiveness figures, usually given to the industry by materials manufacturers, are provided as a guide to show the relative level of shielding performance on an ideal enclosure. That enclosure is one which is solid, without holes, for display, keyboard, access, ventilation, power, or seams, etc. This is not practical. However, properly designed vents can appear to be electrically solid and input/output cables can be appropriately grounded and shielded to minimize their degradation of the equipment's EMI signature when it undergoes EMC acceptance testing.

Flat rectangular panels were evaluated at frequencies from 1 MHz through 10 GHz in a back-to-back screen room test set up similar to that shown in Figure 5. The EMI receiver was enclosed within the confines of a second shielded room to eliminate the possibility of stray radio frequency signals yielding false information. Unshielded plastic, an aluminum reference panel, as well as

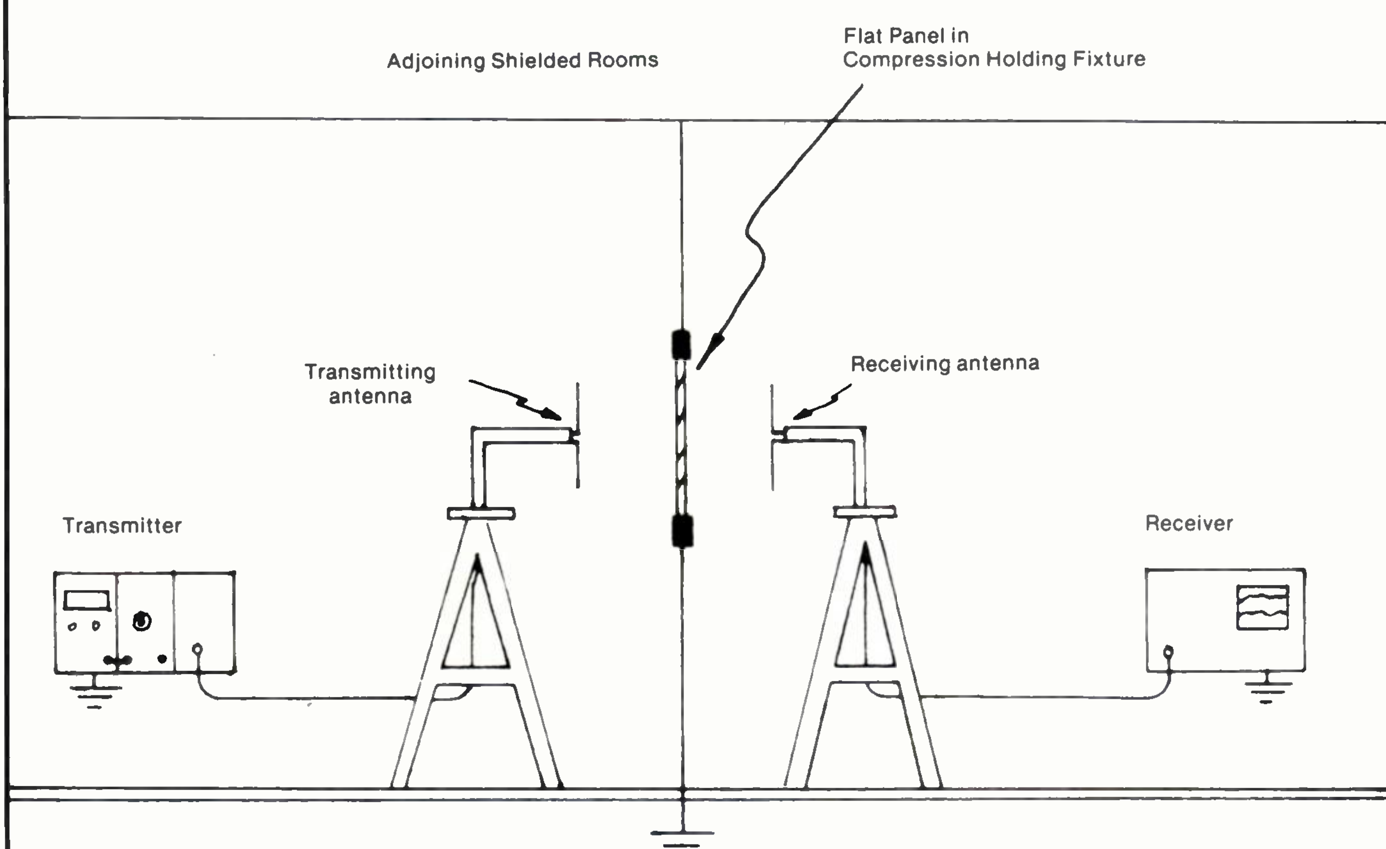

Figure 5. Typical common wall attenuation test set-up.

Table 2.

Material	Thickness Mils	Sheet Resistance Ohms/Square @ 1 mil	Attenuation db
Plastic	1/8 inch	∞	0
Aluminum Sheet	1/8 inch	0	64 - 80
Silver Paint	1.5 mil	0.01	54 - 70
Silver/Graphite (two coat)	0.2/1.0 mil	0.01/100	54 - 77
Copper	2.0 mil	8.0	20 - 54
Copper/Graphite (two coat)	2.0/2.0 mil	8.0/100	27 - 62
Graphite	1.0 mil	100	11 - 60

various conductive approaches* on plastic were tested. Shielding effectiveness variations were noted with changes in frequency and electrical resistance as predicted by microwave theory. As expected, conductivity played a significant role in EMI shielding effectiveness. Table 2 shows the average shielding effectiveness across a range of frequency of 1 MHz to 10 GHz.

*Electrodag®

Testing Methods — ESD

Various plastic-encased microprocessor-based equipments are susceptible to Electrostatic Discharge (ESD) interference. Empirical approaches are tried in order to make a system more immune to ESD problems. An RC circuit consisting of a 500 picofarad capacitor charged to a high voltage in series with a 5000 ohm resistor is used to study ESD in military static sensitive devices.

The level of ESD immunity is found by discharging a high-voltage probe at various points on a typical electronic system housing until the most sensitive area is located. Initially this is done at lower than normal body ESD levels and is continued while gradually increasing the magnitude of the voltage until malfunctions occur. Various methods of grounding and cabling arrangements are tried to minimize the ESD susceptibility. If all else fails, various systems of shielding are employed and tested in the same way until the device has the desired ESD immunity.

An experiment was run to observe the phenomena of charge distribution on high conductivity vs low conductivity shielding materials. Two similar plastic substrates were tested. A coating of highly conductive material was applied to one panel and a low conductivity material to the other (less than 0.1 and greater than 20 ohms, respectively). After scribing through the coatings at about 1/8 inch spacings across the path where current would

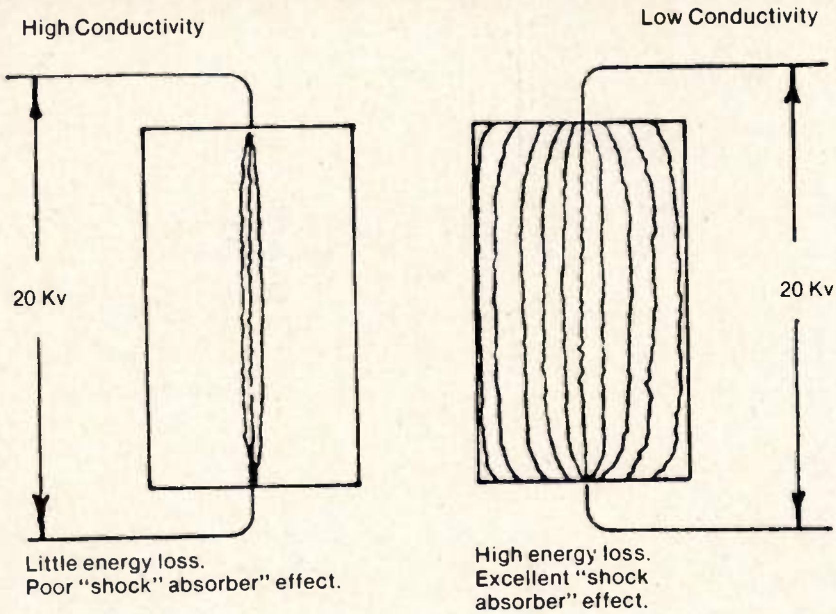

Figure 6. Arc concentration and spread as a function of coating conductivity.

flow, a high voltage source was applied across the panels. The narrow open areas on the surface of the respective conductive materials revealed arc concentration on the highly conductive panel and arc spread on the low conductivity panel. See Figure 6.

Conclusions

1. Consideration of the EMC requirements must address EMI and ESD design goals in shielding product selection. If ESD is the major concern, too conductive a shield may give rise to continued EMI problems.

2. Coatings or metalization methods are only part of the EMC solution. The rest of the solution lies in circuit design, layout, and conventional grounding and filtering considerations which have been in many design arenas left until the system's final days of design — or worse, until after design finalization has taken place and problems occur.
taken place and problems occur.

3. There is no approach which may be assumed to be best, as the best for one system may be enormous over- or underdesign for another system with entirely different susceptibility levels. □

Reference

1. "Limits and Methods of Measurement of Electromagnetic Emanations from Electronic, Data Processing and Office Equipment," CBEMA/ ESC5, 20 May 1977, Washington, D.C.

The FCC has you under the gun.

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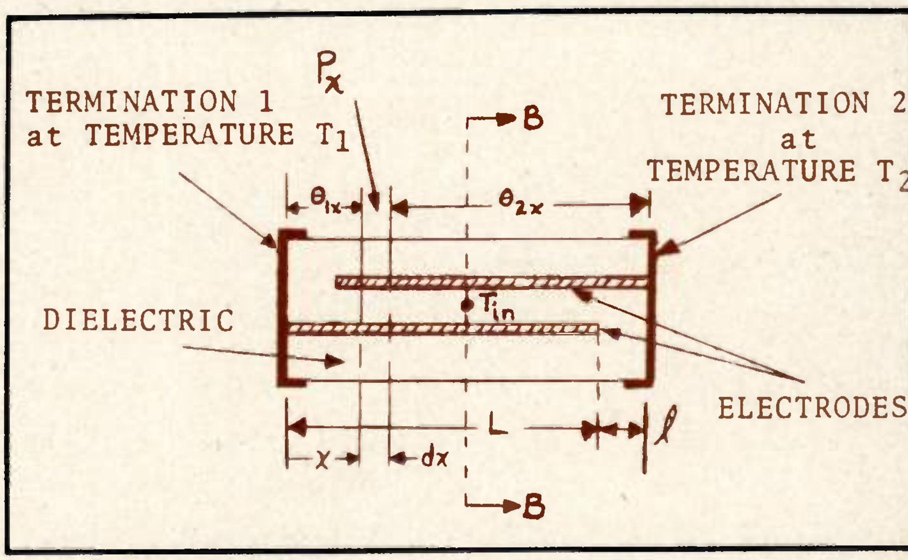
Part I

Ceramic and Porcelain

Thermal Resistance, Power Dissipation and Current Rating.



F.M. Schabauer & R. Blumkin
American Technical Ceramics
Huntington Station, N.Y.

Introduction

The information in this article makes it possible for a circuit designer to calculate the temperature rise of any multilayer capacitor*. The method used for calculation of the temperature rise of a capacitor is quite similar to the techniques that are universally used for transistors.

The theoretical determination of the temperature rise of a capacitor due to AC current flowing through it is a difficult task. Equipment designers, when faced with the problem, require parameters that are generally not available from the capacitor manufacturer, such as ESR (Equivalent Series Resistance), and Θ (Thermal Resistance), etc. of the capacitor.

If the ESR and current are known, the power dissipation and thus, the heat generated in the capacitor can be calculated. From this, plus the thermal resistance of the capacitor and its external connections to a heat sink, it becomes possible to determine the temperature rise above ambient of the capacitor.

Current distribution is not uniform throughout a monolithic capacitor, since the outermost plates (electrodes) carry

*Manufactured by American Technical Ceramics.

Multilayer Capacitors

less current than the inner electrodes. This is shown in Figure 1A for an 8 electrode capacitor. From the figure, it can be seen that there are 7 capacitor sections (Since for N electrodes there are (N-1) capacitor sections.) If the total current into the capacitor is I, the current for each section is I/7. For an outermost electrode, I/7 is actually the current carried by the electrode. For all other electrodes, the current is 2 (I/7) since the electrodes carry the current for two sections. Furthermore, the current is not the same at each point on the electrode. For electrode 8, the current is I/7 at the left or termination end and zero at the right or open end. The current distribution is approximately as shown in Figure 1B. As a result of this current distribution, the heat generated is not uniform within the capacitor.

For an actual multilayer capacitor, there are connection resistances between the electrodes and the terminations, which cause heat generation. This effect depends upon the quality of manufacture of the capacitor. Some manufacturers have fairly high connection resistances, whereas others have connection resistances that are undetectable.

This article assumes a capacitor manufactured with no defects, i.e. zero connection resistances, and it also assumes that the temperature difference across the thickness of the dielectric between electrodes is negligible, i.e. less than 1°C.

The validity of the assumptions has been checked experimentally by measurements of ESR and temperature rise vs. RF current for various capacitor values at a frequency of 30 MHz.

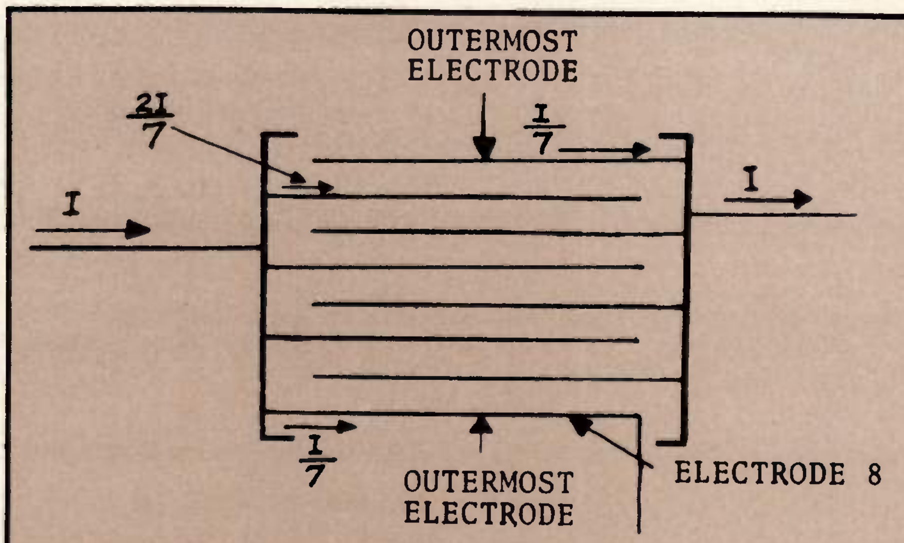


Figure 1A.

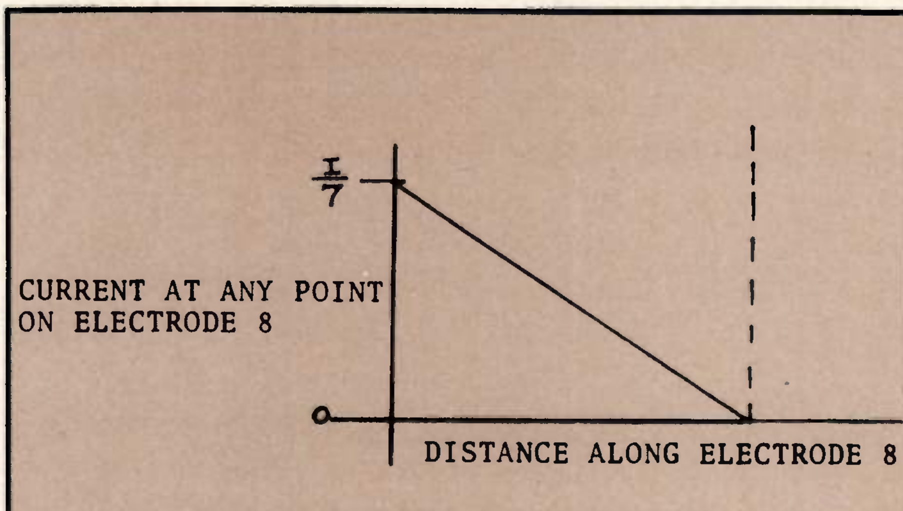


Figure 1B.

Capacitor RF Current Ratings

There are two criteria for maximum current rating.

The first criterion is due to the rated working voltage of the capacitor and is discussed below.

The RF current corresponding to this voltage is:

$$I_p = \frac{V_{peak} - V_{DC}}{X_C} \qquad (1)$$

where,

I_p = Peak RF current

V_{peak} = *Rated Working Voltage* of capacitor

V_{DC} = DC voltage across capacitor

X_C = Reactance of capacitor at frequency of operation

The RF current must not exceed the value from Equation (1).

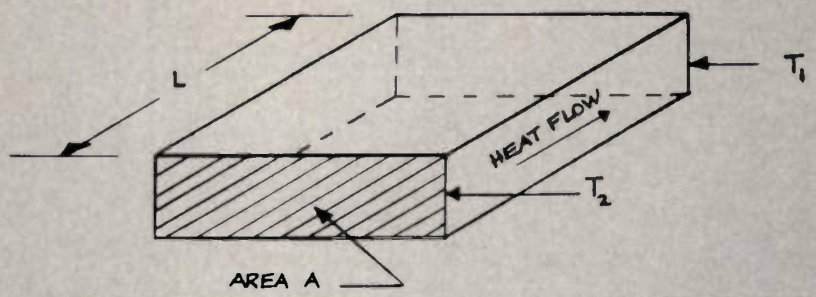
The second criterion is due to the temperature rise caused by power dissipation, (discussed in succeeding paragraphs). In most applications, multilayer capacitors are soldered into the circuit or fastened into place by use of a conductive epoxy. Since the maximum temperature of the solder normally used on the terminations of the capacitor is 190°C, 125°C was chosen as a maximum for one series of capacitors.* This assures the user that the temperature will not exceed the softening temperature of the epoxy or solder. This temperature then determines the maximum power dissipation and in turn, the maximum current, if the capacitor ESR is known.

Working Voltage Rating

The criterion for the maximum voltage rating depends upon the voltage breakdown characteristics of the capacitor. The *voltage breakdown rating* is normally some fraction of the actual internal breakdown voltage. For one series of porcelain dielectric capacitors,** the breakdown voltage exceeds 1000 volts/mil of dielectric thickness and is virtually independent of temperature. Other dielectrics, such as barium titanate and many NPO's have much lower breakdown voltages/mil.

In some situations, the surface breakdown or flash-over voltage rather than the actual internal breakdown voltage is the determining factor. In these cases, the flash-over determines

*ATC 100 series.
**ATC Procelain dielectric capacitors.



where, A = Cross section plane perpendicular to heat flow (cm^2)

P_d = Power dissipated (watts) at area A

T_2 = Temperature (°C) of cross section area A (perpendicular to heat flow)

T_1 = Temperature (°C) at a cross section area at a distance L from area A

L = Length of path (cm) between areas

Θ = Thermal resistance of path across length L (°C/W)

Figure 2.

the rated working voltage. The factors affecting flash-over voltage include surface length of path, surface contamination and environmental conditions.

Current Rating Due To Power Dissipation

Before launching into a thermal analysis of the multilayer capacitor, it is advisable to review some basic thermal principles:

Heat Transfer

The equivalent of Ohm's Law for heat transfer is: (See Figure 2)

$$P_d = \frac{(T_2 - T_1)}{\Theta} \qquad (2)$$

where, P_d is analogous to electrical current,

$(T_2 - T_1)$ is analogous to electrical voltage difference and Θ is analogous to electrical resistance.

Thermal Resistance

The thermal resistance for a given material and dimensions can be calculated:

$$\Theta = \frac{L}{4.186KA} \qquad (3)$$

where,

K = Thermal conductivity coefficient of the material [gm cal/(°C) (sec) (cm)]

L = Length of path (cm)

A = Area perpendicular to path (cm^2)

Note: When the thermal conductivity is given in watts/(°C) (cm), multiply by .2389 to obtain gm cal/(°C) (sec) (cm).

To provide a useful thermal model for calculating the power dissipation of multilayer capacitors, the following constraints are applied:

a) The thermal resistances of the terminations are negligible. This is accomplished by selection of the proper termination material, control of its thickness, uniformity of termination deposition and tight process control.

b) Heat is removed by the conduction mode only, via the terminations of the capacitor to external leads or transmission lines, etc. Radiation and convection are disregarded. This constraint provides an additional safety factor in the current ratings.

c) The thermal conductivity is con-

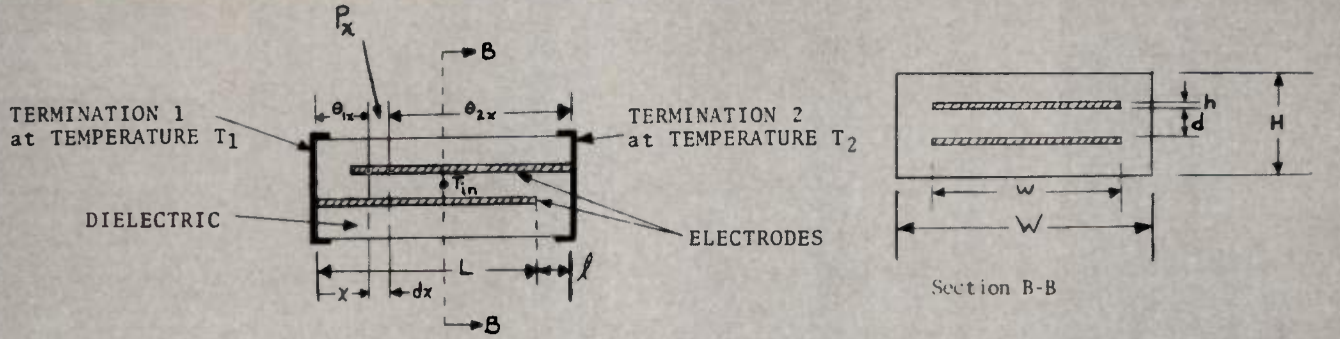


Figure 3.

stant over the temperature range of 25°C to 125°C.

The thermal circuit for a multilayer capacitor is complicated because there are many parallel thermal paths. Since the current varies over the length of the capacitor, the power dissipation is not concentrated at any one point in the capacitor, but is distributed throughout the length of the capacitor. To simplify this situation, an equivalent thermal circuit is derived which substitutes a single lumped power dissipation source (heat generator) at the central plane of the capacitor and a lumped thermal resistance from this central plane to each of the capacitor terminations.

Figure 3 illustrates the deviation of this thermal equivalent circuit for a two electrode capacitor. A strip dx is selected at a distance x from termination 1. The power dissipation in the electrodes in this strip is calculated from i^2R_xdx, where i is the current in one electrode at plane x and R_x is the resistance per unit length of the electrode. Similarly the power dissipation in the dielectric in this strip is calculated from the dissipation factor and the current. The dissipation factor of the dielectric is constant as a function of x. The total power dissipation in the strip dx is P_x and is the sum of the two above power dissipations. The thermal resistances Θ_{1x} and Θ_{2x} from the strip to the terminations consist of parallel electrode and dielectric paths and are calculated from the formulas:

$$\Theta_{1x} = \frac{x}{4.186KA} \quad or \quad \Theta_{2x} = \frac{L-x}{4.186KA}$$

If now the terminations 1 and 2 are connected together thermally but not electrically, i.e. the temperature of termination 1 is the same as the temperature of termination 2, then the temperature rise at plane x of the capacitor can be calculated from the expression:

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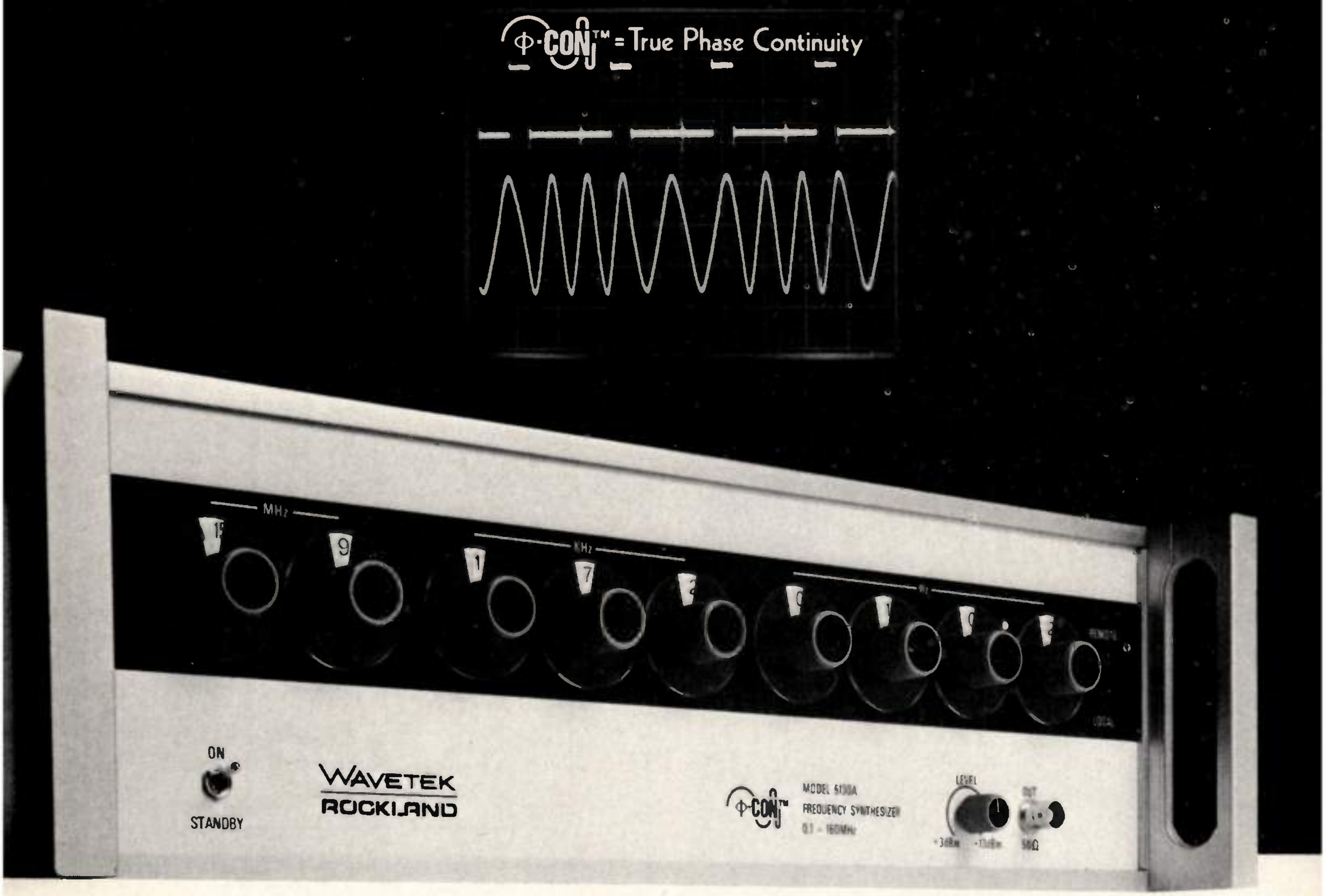

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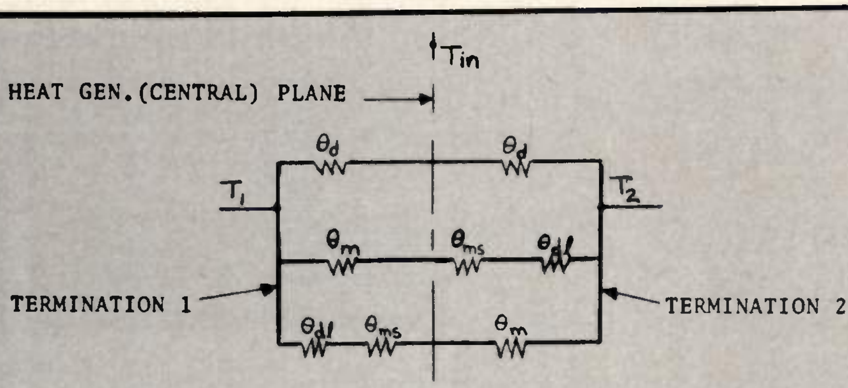

where, θ_d = Thermal resistance of dielectric from Heat Generation Plane to a termination (°C/W)

θ_m = Thermal resistance of parallel combination of all electrodes from Heat Generation Plane to the connected termination for length = $\frac{L+\ell}{2}$ (°C/W)

θ_{ms} = Thermal resistance of parallel combination of all short electrodes from Heat Generation Plane to unconnected end of electrodes for a length = $\frac{L-\ell}{2}$ (°C/W)

$\theta_{d\ell}$ = Thermal resistance of parallel combination of dielectric in series with short electrodes for a length = ℓ (°C/W)

T_{in} = Temperature of Heat Generation Plane (°C)

T_1 = Temperature of termination 1 (°C)

T_2 = Temperature of termination 2 (°C)

Figure 4.

$$\Delta T_x = P_x \frac{\Theta_{1x}\Theta_{2x}}{\Theta_{1x}+\Theta_{2x}}$$

where,

ΔT_x = Temperature rise above T_1 or T_2 (°C)

$\Theta_{1x} = f_1(x)$ = Thermal resistance from plane x to termination 1 (°C/W)

$\Theta_{2x} = f_2(L-x)$ = Thermal resistance from plane x to termination 2 (°C/W)

$P_x = f_3(R_x, x, dx)$ = Power dissipated in metal electrodes and dielectric in width dx located at plane x

If ΔT_x is integrated, an expression is obtained in a form as follows:

$$\Delta T = f\left(P_d, \frac{\Theta}{2}\right)$$

where,

Θ = Thermal resistance from central plane to termination 1 and termination 2 (°C)

P_d = Total Power dissipated in capacitor (watts)

and thus,

ΔT = Temperature rise of central plane above termination (°C)

This permits the establishment of the equivalent circuit with all the power dissipation in the central plane and thermal resistances from that plane to each of the terminations.

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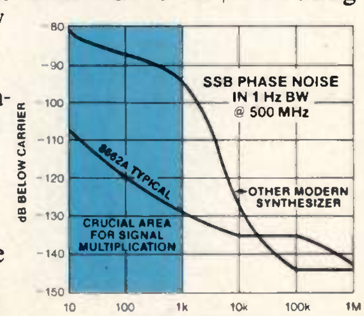

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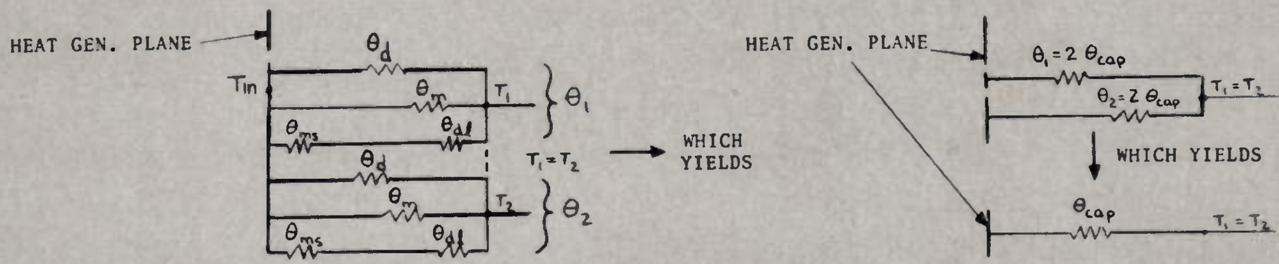

$$\text{where, } \theta_1 = \theta_2 = 2\theta_{cap} = \frac{1}{\left(\frac{1}{\theta_d}\right) + \left(\frac{1}{\theta_m}\right) + \left(\frac{1}{\theta_{ms} + \theta_{d\ell}}\right)} \tag{4}$$

$$\theta_{cap} = \frac{1}{2}\,\frac{\theta_1\,\theta_2}{\theta_1 + \theta_2} \tag{5}$$

θ_{cap} = Thermal resistance of capacitor from Heat Generation Plane to both terminations (°C/W)

Figure 5.

θ_{cap} Calculated From Electrode and Dielectric Dimensions and Thermal Conductivity

	SERIES	100A		100B		
ELECTRODES	Cap Value (pF)	1	100	1	100	1000
	N = Number of Electrodes	2	28	2	18	62
	L (cm)	0.1		0.22		
	ℓ (cm)	0.04		0.06		
	A_m (cm^2)	0.00006		0.000141		
	$N\,A_m$ (cm^2)	0.00012	0.00168	0.000282	0.002533	0.00874
	K_m	0.167 gm cal/(sec)(°C)(cm)				
	θ_m (°C/W)	1670	120	1420	158	46
	θ_{ms} (°C/W)	715	51	812	90	26
DIELECTRIC	$L + \ell$ (cm)	0.14		0.28		
	A_{cap} (cm^2)	0.02		0.07		
	A_d	0.01988	0.0183	0.06972	0.06746	0.06126
	K_d	0.03 gm cal/(sec)(°C)(cm)				
	θ_d (°C/W)	28	30	16	16.5	18
	$\theta_{d\ell}$ (°C/W)	5310	380	3390	376	109
CAP	θ_{cap} (°C/W)	13.7	11.4	7.9	7.2	5.9

subscript d = dielectric subscript m = metal electrode

Equations used in calculation are from equations 3, 4 and 5:

$$\theta_m = \frac{\frac{1}{2}(L+\ell)}{4.186K_m\left(\frac{NA_m}{2}\right)} \qquad \theta_d = \frac{\frac{1}{2}(L+\ell)}{4.186K_dA_d} \qquad \theta_{ms} = \frac{\frac{1}{2}(L-\ell)}{4.186K_m\left(\frac{NA_m}{2}\right)}$$

$$\theta_d = \frac{\ell}{4.136K_d\left(\frac{NA_m}{2}\right)} \qquad \theta_{cap} = \frac{\frac{1}{2}}{\left(\frac{1}{\theta_m}\right) + \left(\frac{1}{\theta_d}\right) + \left(\frac{1}{\theta_{ms} + \theta_{d\ell}}\right)} \qquad \begin{aligned} A_m &= wh \\ A_{cap} &= WH \\ A_d &= WH - whN \end{aligned}$$

NOTE: θ_{cap} PLAYS THE SAME ROLE FOR CAPACITORS AS θ_{jc} PLAYS FOR TRANSISTORS.

Table 1.

The validity of this result is also apparent from the symmetry of the structure of the capacitor on either side of the central plane. This symmetry is also true for the capacitor's power dissipation and thermal resistances.

Figure 4 is the thermal equivalent circuit for the two electrode capacitor in Figure 3. From Figure 4 one can see that there are two equal thermal paths from the central plane to each of the terminations. For *each* path there are *three* thermal resistances in parallel. One is metal, the second is dielectric and the third is metal in series with a small length (ℓ) of dielectric. The first and third are through the cross-sectional area of the electrodes (wh) and the other is through the area of the dielectric (WH-2wh). If there are N electrodes, these become Nwh/2 and (WH-Nwh).

If termination 1 is thermally connected, but not necessarily electrically connected to termination 2, T_1 becomes equal to T_2. This is equivalent to folding Figure 4 at the Heat Generation Plane and connecting termination 1 to termination 2.

The thermal resistance of the capacitor is thus developed as shown in Figure 5.

Using the equivalent circuit of Figure 5 and equations 3, 4 and 5 the thermal resistance of ATC 100A 1pF and 100pF capacitors and ATC 100B 1pF, 100pF and 1000pF capacitors can be calculated. The results are shown in Table 1. □

Part 2 will appear in the next issue of *r.f. design*.

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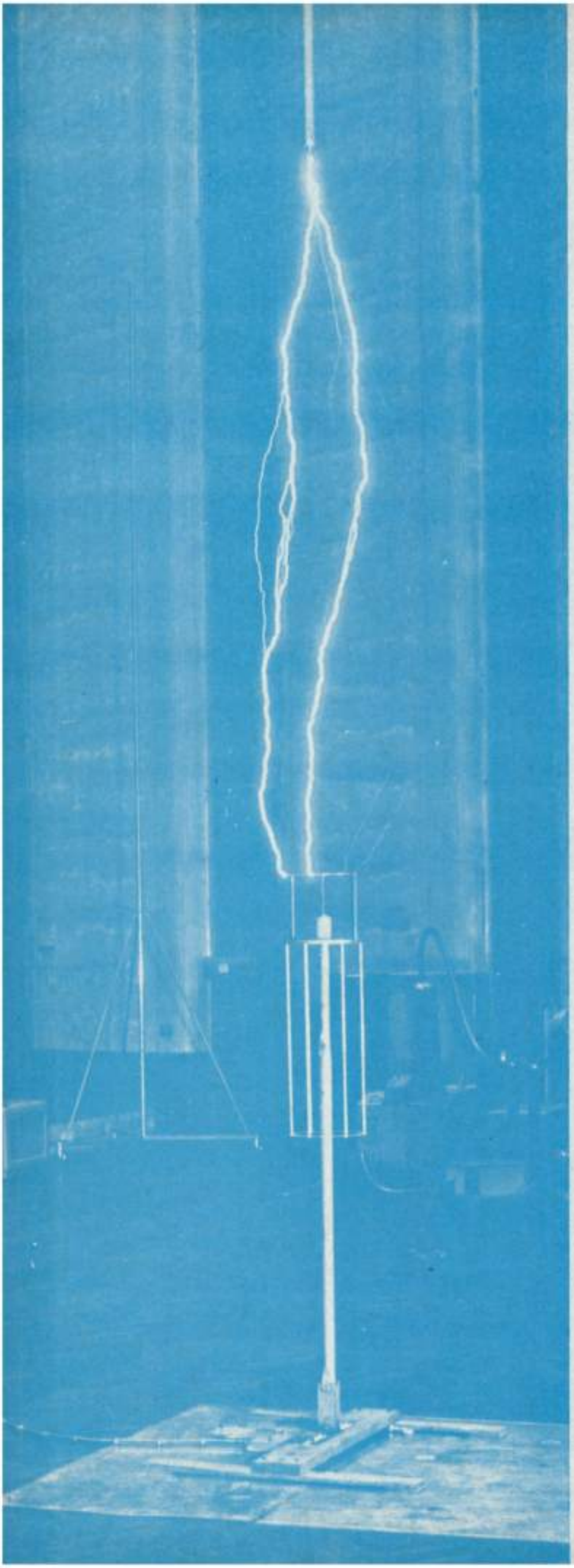

Active

A description and system analysis of an antenna that is an integrated combination of a short passive element and gain amplifier.

Ulrich L. Rohde, Ph.D.
Communications Product Corp.
Upper Saddle River, NJ

The active antenna in its minimum configuration consists of a passive antenna, typically, a rod or a dipole and an integrated amplifying device.

Let us look at the simple case in which a rod antenna is directly connected to the input of a field effect transistor. As shown in Figure 1, the antenna acts as a source that feeds the transistor. The electric field strength E generates a voltage (EMF) that can be determined from $V_a = E \cdot h_e$. The antenna has a capacitance C_A and for small electrical lengths, this is 25 pFd/meter, while the transistor has an input capacitance C_T. These two capacitances form a capacitive voltage divider. The signal voltage that drives the transistor is then

$$V_T = \frac{E \cdot h_e}{1 + C_T/C_A}$$

For electrically short antennas the voltage V_T is nearly independent of frequency. Therefore, the active antenna has an extremely wide bandwidth.

The gain-bandwidth product of such a device can be computed from the performance of the field effect transistor in Figure 1. It will reproduce at the output the input voltage as long as its cut-off frequency is high enough. Additional reactances (for frequency selectivity) may be added to intentionally limit the bandwidth of the active antenna.

Antennas

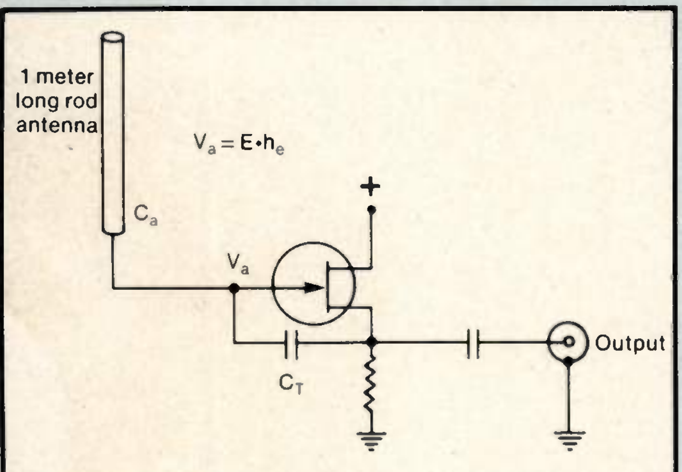

Figure 1.

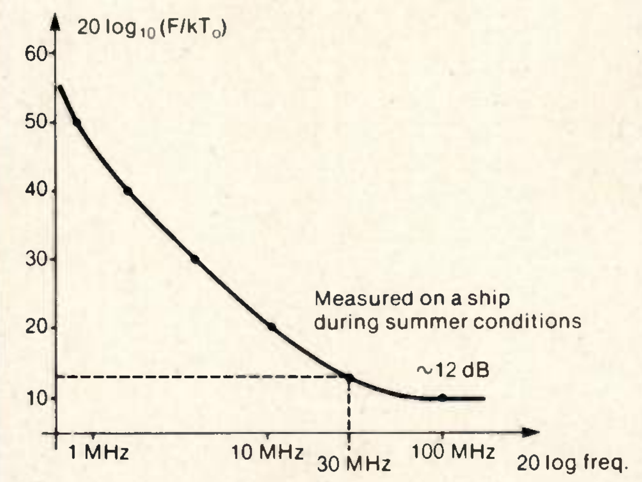

Figure 2. Atmospheric noise as a function of frequency.

Output power is not considered of primary importance since post amplifiers can always be added. Therefore, only the signal-to-noise ratio is worth considering.

Assume that the active antenna has sufficient gain. Then the signal-to-noise ratio is determined by the active antenna and not by the receiver. The only internally-generated noise is from the transistor since the passive antenna must be considered noise free.

In analyzing the output of the active antenna, there are three components to consider:

a) The signal voltage at the operating frequency,

b) The amplified noise generated by external sources (man-made or galactic), and

c) The transistor noise contribution.

As long as the noise voltage generated by the transistor at the amplifier output is less than the wideband noise picked up by the antenna, the system is capable of supplying the same signal-to-noise ratio as an optimized passive antenna for the same specific frequency.

Let us assume for a moment that we have an active antenna with a 1 meter long rod element. Its capacitance is 25 pFd. A typical value for the FET is 5 pFd or one fifth. Consequently 80 percent of the antenna output is applied to the FET input.

If a passive full wavelength long dipole were used instead at, say, 10 MHz (30 meters long), the open-circuit voltage (EMF) would be 30 times higher than that generated by the 1 meter long rod. In addition, the atmospheric noise term would be greater. However, the 1 meter rod, for all practical purposes, generates the same, or practically the same, signal-to-noise ratio as a dipole with the difference that individual (signal and noise) levels are smaller. The difference in amplitude can be compensated for by an amplifier under the restriction that atmospheric noise divided by the ratio of (full size antenna/1 meter) is equal to or better than the noise figure of the transistor amplifier.

The average noise power of the ionosphere at these low frequencies is seen in Figure 2. These are the figures measured in typical rural areas. It becomes apparent that if these voltages are divided by 50, the noise floor approaches the noise floor of the best FETS, i.e. approximately 1 to 2 dB.

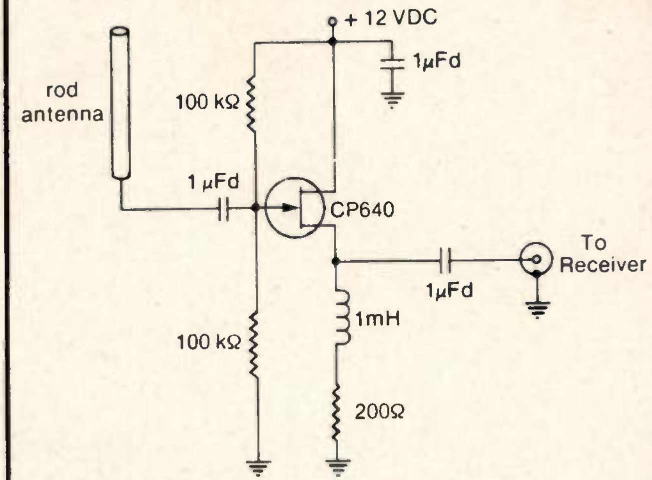

Figure 3.

Intermodulation Distortion

Intermodulation distortion is another type of noise that is generated in the transistor or active device as a function of the input signals. The active antenna, therefore, is best described by assigning to it:

a) A frequency range
b) A minimum sensitivity which is determined by the noise figure
c) A dynamic range which is determined by the second, third, and higher-order intercept points, and
d) Polarization (horizontal or vertical).

The simplest active antenna configuration using an FET is illustrated in Figure 3. This circuit exhibits high second-order intermodulation distortion due to the antenna square-law characteristic. A push-pull configuration that can be used for tests and evaluations of such a system is illustrated in Figure 4.

Table 1. Active Antenna Dynamic Analysis Terminology.

a	Cable losses
B	Receiver bandwidth
C	Noise correlation factor
F_A	Antenna noise figure
F_{min}	Amplifier noise figure (for best noise match)
F_R	Receiver noise figure
F_S	System noise figure (antenna, cable, and receiver)
G_v	Antenna gain = antenna output power
IP_2	Second-order intercept point
IP_3	Third-order intercept point
P_a	Output power into 50 ohms
$P_{am_{2,3}}$	Second or third order intermodulation products output power
P_{an}	Noise output power
V_a	Output voltage (terminated in 50 Ω)
V_o	Antenna output EMF
Z_A	Antenna rod impedance
Z_{opt}	Antenna rod impedance (for best noise matching)

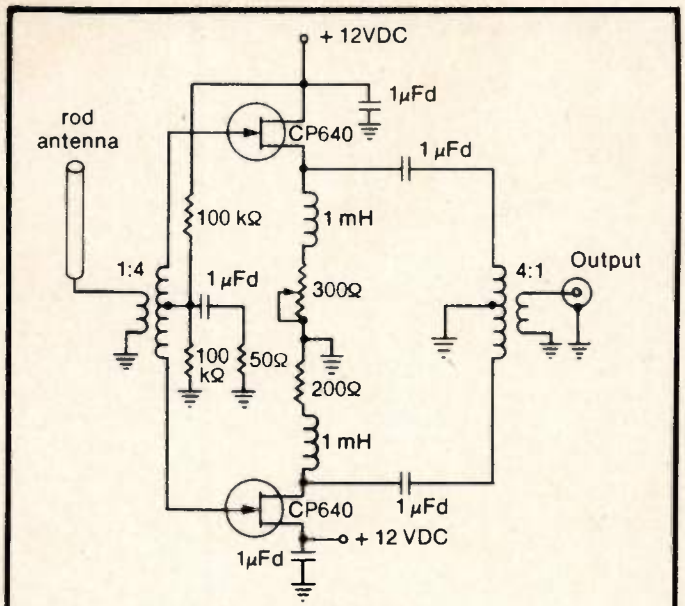

Figure 4.

Active Antenna Dynamic Analysis
(See Table 1 for terminology.)

System noise figure is defined by

$$F_S = F_A + \frac{(F_R - 1) \cdot a}{G_v}$$

The electrical gain of the antenna is defined by

$$G_v = 4\left(\frac{V_a}{V_o}\right)^2 \cdot \frac{R_A}{Z_L}$$

For reasons of best dynamics, assume $V_a/V_o = 0.5$.

With these assumptions in mind, the noise figure of the antenna now becomes

$$F_A = F_{min}\left(1 + C\,\frac{(Z_A - Z_{opt})^2}{R_A \cdot R_{opt}}\right) = F_{min}(1 + A)$$

The impedance Z_A and Z_{opt} are

$$Z_A = R_A + jX_A$$

$$0.25 < C < 0.5$$

$$Z_{opt} = R_{opt} + jX_{opt}$$

If the antenna is set for the greatest possible bandwidth, X_{opt} becomes 0. The antenna noise figure then is

$$F_A = C\left(\frac{R_A}{R_{opt}} + \frac{R_{opt}}{R_A} + \frac{X_A^2}{R_A \cdot R_{opt}} - 2\right)$$

This particular type of matching requires a high input impedance. Therefore:

$$\frac{R_A}{R_{opt}} << \frac{R_{opt}}{R_A}, \quad \frac{R_A}{R_{opt}} + \frac{R_{opt}}{R_A} + \frac{X_A^2}{R_A \cdot R_{opt}} >> 2$$

Table 2.

f (MHz)	G_V (dB)	$F_A \left(\frac{dB}{kT_o}\right) \approx F_S \left(\frac{dB}{kT_o}\right)$
2	−31	47.4
5	−21.7	35.2
10	−15.5	28.6
15	−12	24.9
20	−10.2	23.1
25	− 7.3	20.2
30	− 5.5	18.4

Finally, the antenna noise figure is

$$F_A = F_{min} \left\{ 1 + \frac{C}{R_A} \left(R_{opt} + \frac{X_A^2}{R_{opt}} \right) \right\}$$

Using a rod antenna, its impedance is

$$Z_A \approx K_{R_1} \omega^2 + j \frac{k_{X_1}}{\omega}$$

The impedance diminishes much faster than the noise figure does (as a function of frequency). Consequently, optimum matching resistance should be specified at the lowest operating frequency. Consider a 2-30 MHz active antenna. Its match resistance is 2466Ω (at 2 MHz). The antenna performance can now be determined if Z_A is known.

Loss and Noise Figure Versus Frequency

Table 2 lists electronic losses and noise figure as a function of frequency. (This assumes that the noise figure of the active device is 2 dB). This data is also plotted in Figure 5. In this graph, the system's noise figure, the man-made noise, and some arbitrary noise specifications are plotted.

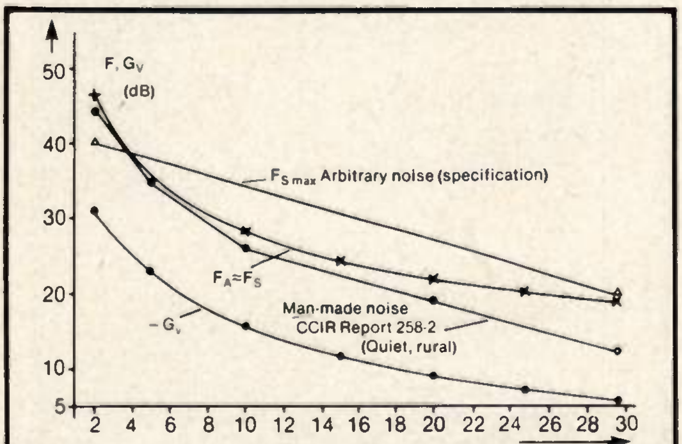

Figure 5. Frequency (MHz)

Despite a loss of 30 dB (the active antenna relative to the power available at the antenna input) the signal-to-noise ratio up to 4 MHz can meet the specifications. Below 4 MHz, the specifications are equal to the man-made noise. Above 4 MHz the antenna's performance exceeds the specifications.

Shipboard Environment Specifications

The active antenna, per the specifications,* sees two 10V EMF's. The intermodulation distortion products that are generated due to these two voltages are 40 dB above the specified maximum system's noise figure, as a worst case condition. Therefore:

$$\begin{aligned} P_{an}(dBm) &= F_S(dB) + G_v(dB) = 10 \cdot \log kToB \cdot 10^3 \\ &= F_S(dB) = G_v(dB) - 139\,dBm \text{ and} \\ P_{am_{2,3\,max}}(dB) &= P_{an\,min}(dBm) + 40\,dB \\ &= [F_S(dB) + G_v(dB)]_{min} - 99\,dBm \end{aligned}$$

At 2 MHz, F_S equals 40 dB and G_v equals −31 dB. Therefore, $P_{am_{2,3}}$ = −90 dBm.

Intercept Point Calculations

$$IP_2(dBm) = 2\,P_a(dBm) - P_{am\,2}(dBm)$$

$$P_a = \frac{V_a^2}{50\Omega}$$

With V_a/V_o = 0.5 and V_o = 10V, P_a is +27 dBm, and, therefore, IP_2 = 144 dBm and IP_3 = 85 dBm. These are the two values that are required to generate an intermodulation distortion noise floor at the rated level. For practical considerations the 1 dB compression point should be 10 dB above the operating output level. Therefore, in this case, it should be +37 dBm. This results in a voltage level of 44.3V at 0.9A in a 50 ohm system. The operating voltage of this amplifier should be set at 50V. If the input voltage ratio is changed and a higher than 0.5 voltage division ratio is utilized, then the second and third order intercept points can be reduced. Let us assume that an intercept point of IP_2 = 100 dBm and IP_3 of 65 dBm can be reached in a practical amplifier. The following results will then be obtained:

1. Second order intermodulation distortion products are going to be −46 dBm and the useful dynamic range will be *84 dB*.
2. Third order intermodulation products will be −49 dBm and the useful dynamic range will be *81 dB*.

These calculations assume a noise figure of 40 dB at 2 MHz and two 10V random carriers generating the intermodulation distortion products as specified.

A number of tests in extremely hostile environments have already been performed with this active antenna. However, it is not yet in mass production and, therefore, not enough information about reproducibility is available. This will be the next step for evaluation.

*Antenna system developed for use on shipboard by Communications Consulting Corp., Upper Saddle River, NJ, based on some discussions with the Naval Research Laboratory in Washington, D.C.

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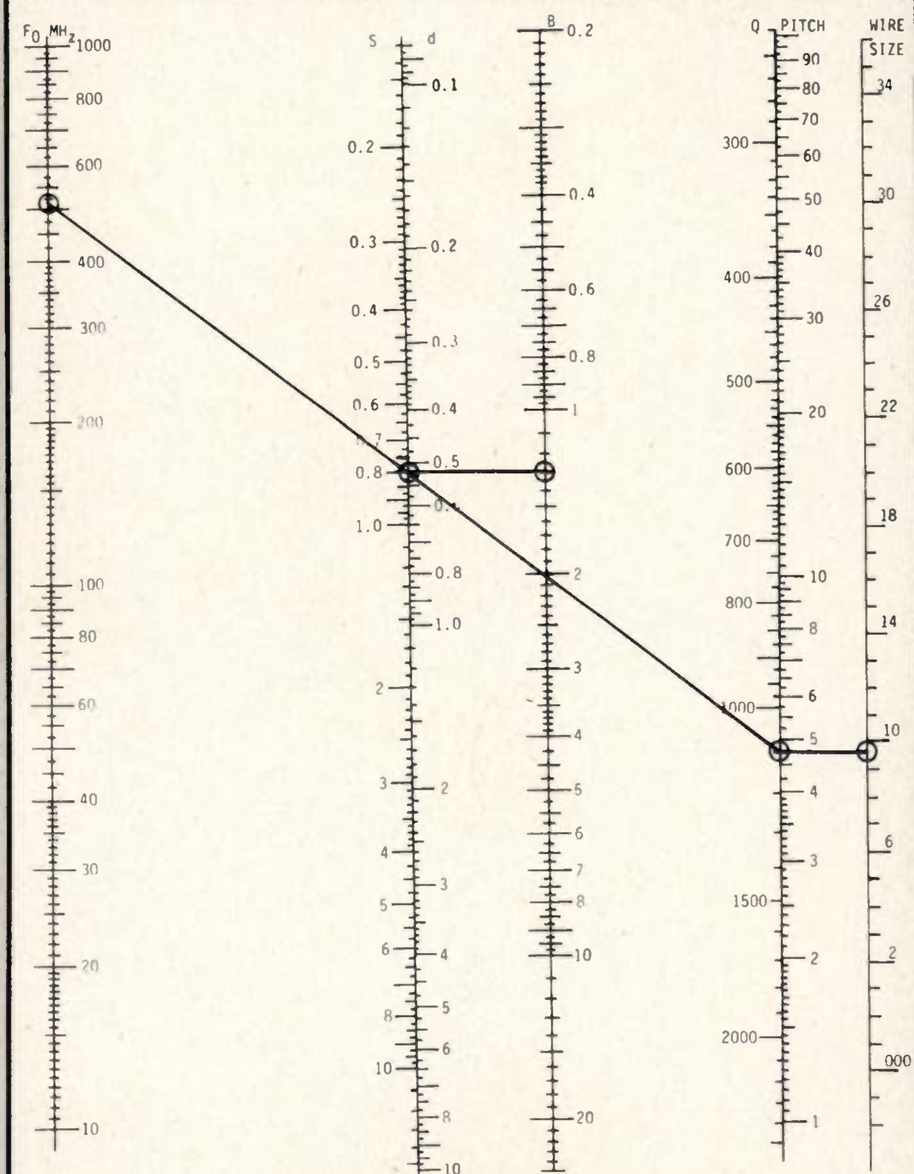

Figure 1. Helical Filter Nomograph.

Need a Helical Filter?

These Design Aids Will Make it Easy

Lee R. Watkins
Martin Marietta
Waterton, CO

Today the design engineer is called upon to generate designs that are at or near the state-of-the-art. But with the amount of knowledge and the rate that it is advancing, it is impossible for any designer to be an expert in very many fields. Most engineers do not consider themselves as experts in the area of filter design. Therefore, when their filter requirements exceed their basic knowledge of filters, they turn to others in the field. This results in increased delays and costs.

Filters in the frequency range from 30 MHz to 1000 MHz often present special problems. *Crystal filters* are possible only in the lower frequencies and for bandwidths in the order of 1 percent or less. *Narrowband low-loss discrete filters* are often difficult or impossible to build because of the high component Q required. Also with low element Q the reject skirts of the filter can be seriously altered. *Coaxial resonator filters* have high Q but at these frequencies become excessively large. (A 1/8 wave resonator at 30 MHz is approximately 50 inches long.) *Stripline filters* are also large and usually suffer from low Q. *Surface Acoustic Wave filters* are being developed and are available for much of this frequency range, but they have high insertion loss, poor dynamic range, limited out-of-band rejection, and limited power handling capability.

The *helical filter* presents a workable solution to the problem of filters in this frequency range. This paper condenses the necessary equations and design aids so that the design engineer can design helical filters with ease and confidence.

The Design

The design procedure is illustrated by an example.

Filter Requirements (Step 1)

The first step is to specify the filter parameters.

Center frequency	520 MHz
3 dB bandwidth	10 MHz
50 dB bandwidth less than	90 MHz
Source and load impedance	50 ohm
Maximum allowable insertion loss	1.0 dB

From data curves[1,2,3] it is determined that a 3rd order Butterworth filter will meet the selectivity requirements. quirements.

Determining Resonator Q Requirements (Step 2)

The insertion loss of a filter is directly related to the ratio of the element Q, (Q_E) and the filter Q, (Q_F). Exact formulas exist but are cumbersome and difficult to work with.[2,4] An approximate formula exists which provides very good accuracy for Q ratios (Q_E/Q_F) greater than 5. This approximate equation is:

$$\text{Insertion loss (I.L.)} = \frac{4.343\, Q_F}{Q_E} \cdot \sum_{K=1}^{N} X_K$$

where:
N is the order of the filter
Q_E is the element Q or unloaded resonator Q
Q_F is the filter Q and is given by $f_0/BW_{3\,dB}$
X_K is the normalized lowpass element values for the given filter type used. See Tables I through VIII.

Since we wish to determine the required resonator Q, (Q_E) that provides an insertion loss of less than 1.0 dB, solve the above equation for Q_E.

$$Q_E = \frac{(4.343)\, Q_F}{I.L.} \sum_{K=1}^{3} X_K$$

$$X_1 = 1.00,\ X_2 = 2.00,\ X_3 = 1.00$$

$$Q_E = \frac{(4.343)(520 \times 10^6)}{(1.0)(10 \times 10^6)} \cdot (4) = 903$$

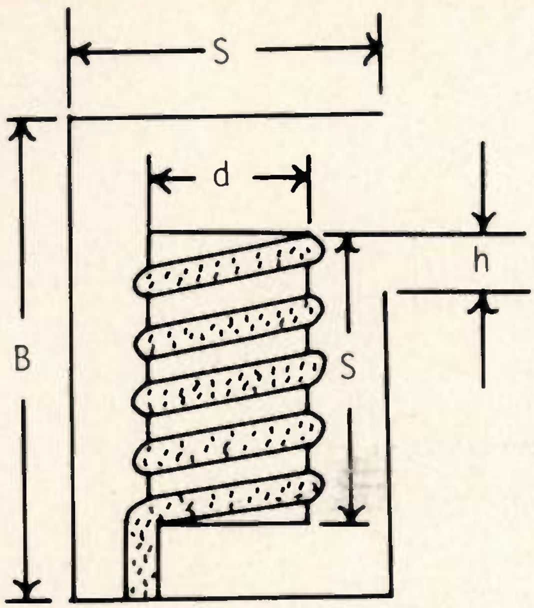

Figure 2. Resonator and Cavity Dimensions.

This value represents the theoretical minimum Q that will give an insertion loss of 1.0 dB. Since we wish to assure a maximum insertion loss of 1.0 dB, it is necessary to require a resonator Q, (Q_E) which is higher than 903. Therefore, we shall arbitrarily choose as a requirement for the resonator Q a value of 1100.

The Physical Dimensions (Step 3)

At this point in the design the physical dimensions of the resonator and cavity are obtained. The dimensions are for optimum Q for the sizes given. All dimensions are given in inches. By using the nomograph of Figure 1 we can find the dimensions for Figure 2.

The center frequency and the required value of Q_E are known. A straight line is drawn between the point representing the center frequency and the point representing the desired Q_E. A horizontal line is then drawn from the intersection of the Line S to the Line B and a horizontal line is drawn from the Line Q to the line labeled wire size.

The values of S, d, B, Pitch, and wire size are then read

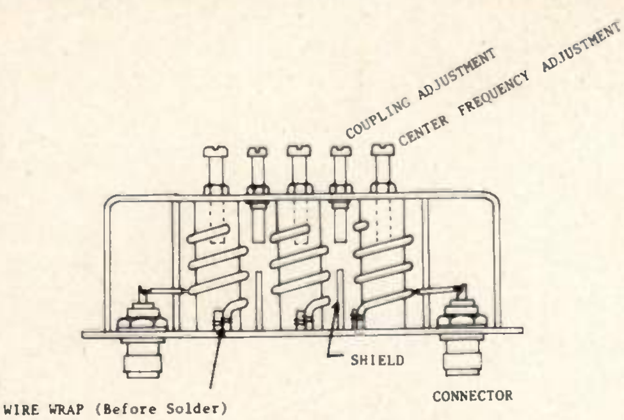

Figure 3. Typical Helical Filter Construction.

from the nomograph. For this case they are:

S = 0.8 inches
d = 0.52 inches
B = 1.30 inches
Pitch = 5 turns/inch
Wire Size = #10

The dimensions S and B are the inside dimensions of the cavity.

The Coupling Dimensions (Step 4)

The dimension h in Figure 2 is the height of the opening from the top of the shield to the top of the coil. Actually the opening extends to the top of the cavity, but the value computed is to the coil height.

The value of h is in general different between each resonator and is a function of the value of d and the value of the normalized lowpass element values (See Table I through VIII).

The equation for h is:

$$h_{ij} = d\,(1.075)\,10^{\frac{1}{1.91}\log_{10}\left(\frac{BW_{3\,dB}}{(0.071)(f_0)(\sqrt{X_i X_j})}\right)}$$

For this case $d = 0.52$

$$X_1 = 1 \quad X_2 = 2 \quad X_3 = 1$$

$$h_{1,2} = h_{2,3} = 0.235 \text{ inches}$$

In finding $h_{1,2}$ which is the coupling height between resonator #1 and resonator #2, the lowpass element values X_1 and X_2 are used. This equation is valid (within about 6 percent) for a shield thickness of 1/16 inch. For a shield thickness of 1/32 inch divide by 1.075.

Locating the Tap Point (Step 5)

In order to match the filter to a 50 ohm system it is necessary to tap into the first and last resonator. The tap position is found by:

Tap point (in turns above ground):

$$T.P. = \frac{N\Theta}{90} = 0.15$$

where:

N = *total number of turns on the resonator.*

$$\Theta = \text{Arcsin}\left(\frac{R_b R_{tap}}{2 Z_0^2}\right)^{1/2}$$

R_{tap} = *50 ohms (system impedance)*

$$R_b = \frac{\pi Z_0}{4} \cdot \left(\frac{1}{Q_d} - \frac{1}{Q_R}\right)$$

$$Z_0 = \frac{8.51 \times 10^{10}}{f_0 S} \text{ (S is found in Step 3)}$$

$$Q_d = \frac{f_0 X_1}{2\,BW_{3\,dB}}$$

Q_R = *1100 (found in Step 2)*

In practice the tap point will need to be adjusted slightly for the best match.

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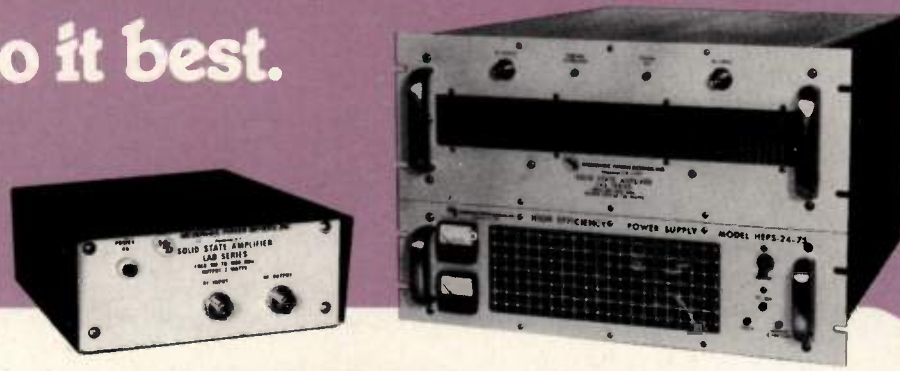

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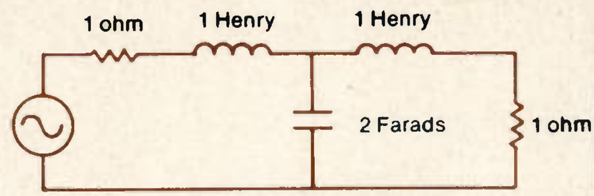


3rd order Lowpass Filter

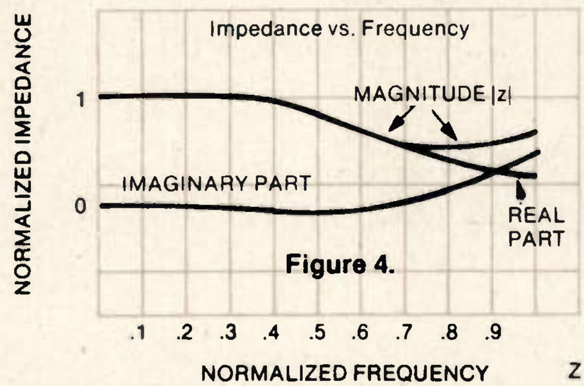


Figure 4.

Tap Adjustment (Optional) (Step 6)

In step five we found the required tap location to be 0.15 turns above ground. For critical designs the tap position must be changed slightly (because the 0.15 location cannot be precisely located) in order to provide the proper impedance. This process can be very tedious and often difficult to do. An alternative method exists which adjusts the tap location by using an impedance transformation. If this method is used there will be some additional loss introduced into the filter of about 0.1 to 0.2 dB. Therefore, this needs to be considered in the overall design requirements.

Figure 4 shows the normalized impedance characteristics of a 3rd order Butterworth filter as a function of frequency. If such a filter were placed on the input and output of the Helical filter and designed so that the cutoff occurred at $f_0/.6$, then the impedance could be tuned by adjusting

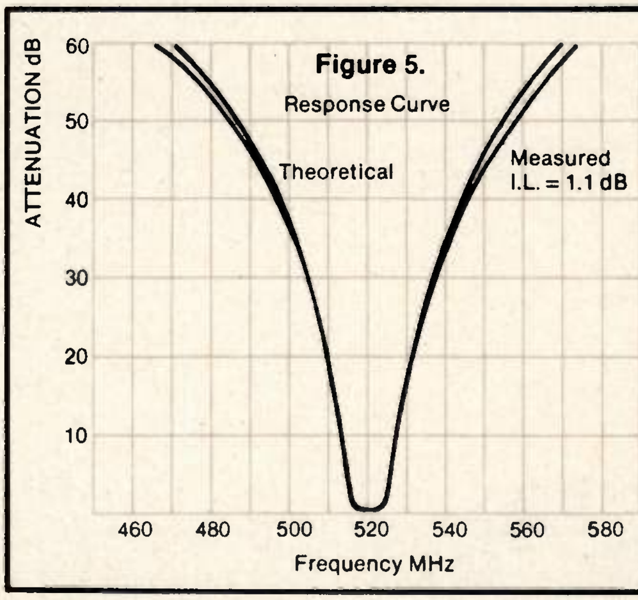
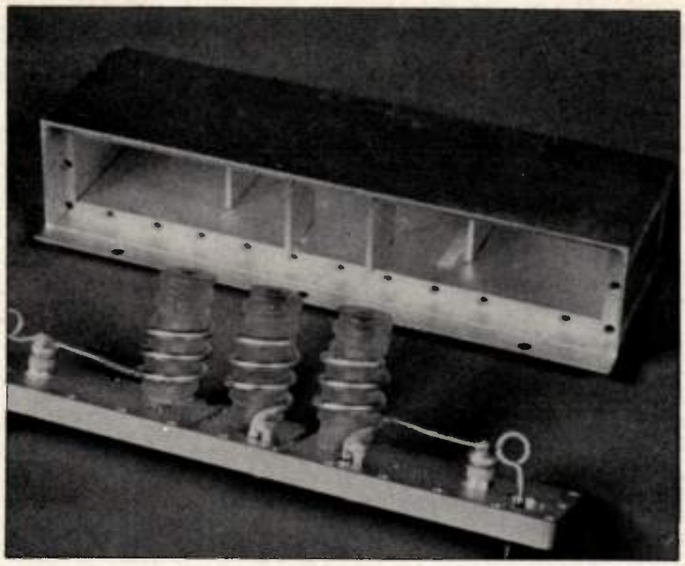


Figure 5.
Response Curve



the trimming capacitor. This method was employed in this example.

At this point the overall measurements of the filter can be specified. The *width* is the dimension S plus the wall thicknesses. The *height* is the dimension B plus the cover and base thicknesses. The *length* is the number of resonators times S plus all shield and end-wall thicknesses.

Input/output connectors will add to the overall length. A sketch of the filter is shown in Figure 3. The resonators are best supported with a coil form as shown. Care must be used in order not to degrade the resonator Q. Experience has shown that either Polystyrene or Rexolite may be used.

When making the housing a number of assembly techniques can be used; dip braze, casting, machining, or piece part assembly. Any method may be used so long as all joints provide a very good junction. This is extremely important for junctions near the base of the resonators. The resonators themselves must be soldered or brazed to the base to insure good contact.

After assembly, filter tuning is accomplished by frequency sweeping the filter and looking at the detected output and adjusting the tuning screws. The screws over the resonators adjust the center frequency and the screws between adjacent resonators adjust the coupling. (Coupling is related to the bandwidth of the filter). For a more thorough treatment of the helical filter tuning see Reference 2. The completed filter along with the response curve is shown in Figures 5 and 6.

To many the helical filter is a mystery. But they are easy to build and except for the most demanding of applications will not cause problems in construction or tuning. It will be found that if these design aids are followed a successful helical filter can be designed, assembled, and tuned.

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TABLE I. Butterworth Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	1.4142	1.4142								
3	1.0000	2.0000	1.0000							
4	0.7654	1.8478	1.8478	0.7654						
5	0.6180	1.6180	2.0000	1.6180	0.6180					
6	0.5176	1.4142	1.9319	1.9319	1.4142	0.5176				
7	0.4450	1.2470	1.8019	2.0000	1.8019	1.2470	0.4450			
8	0.3902	1.1111	1.6629	1.9616	1.9616	1.6629	1.1111	0.3902		
9	0.3473	1.0000	1.5321	1.8794	2.0000	1.8794	1.5321	1.0000	0.3473	
10	0.3129	0.9080	1.4142	1.7820	1.9754	1.9754	1.7820	1.4142	0.9080	0.3129

TABLE II. Chebyshev 0.01 dB Ripple Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	1.3472	1.4830								
3	1.1811	1.8214	1.1811							
4	0.9500	1.9382	1.7608	1.0457						
5	0.9766	1.6849	2.0367	1.6849	0.9766					
6	0.8514	1.7960	1.8411	2.0266	1.6312	0.9372				
7	0.9127	1.5947	2.0021	1.8704	2.0021	1.5947	0.9127			
8	0.8145	1.7275	1.7984	2.0579	1.8695	1.9796	1.5694	0.8966		
9	0.8854	1.5513	1.9615	1.8616	2.0717	1.8616	1.9615	1.5513	0.8854	
10	0.7970	1.6930	1.7690	2.0395	1.8827	2.0724	1.8529	1.9472	1.5380	0.8773

TABLE III. Chebyshev 0.1 dB Ripple Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	1.2087	1.6383								
3	1.4329	1.5937	1.4329							
4	0.9924	2.1476	1.5845	1.3451						
5	1.3013	1.5559	2.2411	1.5559	1.3013					
6	0.9419	2.0798	1.6581	2.2473	1.5344	1.2767				
7	1.2615	1.5196	2.2393	1.6804	2.2393	1.5196	1.2615			
8	0.9234	2.0455	1.6453	2.2826	1.6843	2.2300	1.5091	1.2516		
9	1.2447	1.5017	2.2220	1.6829	2.2957	1.6829	2.2220	1.5017	1.2447	
10	0.9147	2.0279	1.6346	2.2777	1.6962	2.2991	1.6805	2.2155	1.4962	1.2397

TABLE IV. Chebyshev 0.3 dB Ripple Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	1.0710	1.8171								
3	1.6854	1.3985	1.6854							
4	0.9601	2.3969	1.4127	1.6290						
5	1.6010	1.4039	2.4956	1.4039	1.6010					
6	0.9343	2.3676	1.4802	2.5116	1.3954	1.5853				
7	1.5756	1.3891	2.5116	1.4991	2.5116	1.3891	1.5756			
8	0.9248	2.3493	1.4784	2.5523	1.5042	2.5084	1.3846	1.5692		
9	1.5648	1.3813	2.5049	1.5053	2.5662	1.5053	2.5049	1.3813	1.5648	
10	0.9204	2.3395	1.4745	2.5537	1.5154	2.5713	1.5050	2.5018	1.3788	1.5617

TABLE V. Chebyshev 0.5 dB Ripple Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	0.9827	1.9497								
3	1.8637	1.2804	1.8637							
4	0.9202	2.5865	1.3036	1.8259						
5	1.8069	1.3025	2.6915	1.3025	1.8069					
6	0.9053	2.5775	1.3675	2.7134	1.2990	1.7962				
7	1.7896	1.2961	2.7177	1.3848	2.7177	1.2961	1.7896			
8	0.8998	2.5671	1.3697	2.7585	1.3903	2.7176	1.2938	1.7853		
9	1.7823	1.2921	2.7163	1.3922	2.7734	1.3921	2.7163	1.2921	1.7823	
10	0.8972	2.5611	1.3683	2.7632	1.4009	2.7796	1.3927	2.7148	1.2908	1.7801

TABLE VI. Chebyshev 1.0 dB Ripple Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	0.8341	2.2185								
3	2.2157	1.0884	2.2157							
4	0.8310	2.9813	1.1208	2.2104						
5	2.2072	1.1280	3.1025	1.1280	2.2072					
6	0.8291	3.0056	1.1788	3.1353	1.1300	2.2052				
7	2.2039	1.1306	3.1470	1.1937	3.1470	1.1306	2.2039			
8	0.8283	3.0077	1.1849	3.1903	1.1994	3.1518	1.1308	2.2031		
9	2.2025	1.1308	3.1540	1.2020	3.2077	1.2020	3.1540	1.1308	2.2025	
10	0.8279	3.0076	1.1862	3.2006	1.2091	3.2159	1.2033	3.1550	1.1307	2.2020

TABLE VII. Gaussian Magnitude Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	2.1850	0.4738								
3	2.2262	0.8167	0.2624							
4	2.2450	0.9321	0.5302	0.1772						
5	2.2533	0.9782	0.6485	0.3896	0.1312					
6	2.2568	0.9982	0.7050	0.5004	0.3045	0.1026				
7	2.2583	1.0073	0.7333	0.5606	0.4055	0.2473	0.0833			
8	2.2590	1.0116	0.7479	0.5942	0.4658	0.3388	0.2065	0.0695		
9	2.2593	1.0137	0.7556	0.6134	0.5025	0.3973	0.2892	0.1761	0.0591	
10	2.2594	1.0147	0.7597	0.6244	0.5250	0.4353	0.3451	0.2509	0.1525	0.0512

TABLE VIII. Maximally Flat Group-Delay Normalized Element Values.

ORDER	X_1	X_2	X_3	X_4	X_5	X_6	X_7	X_8	X_9	X_{10}
2	2.1478	0.5755								
3	2.2034	0.9705	0.3374							
4	2.2404	1.0815	0.6725	0.2334						
5	2.2582	1.1110	0.8040	0.5072	0.1743					
6	2.2645	1.1126	0.8538	0.6392	0.4002	0.1365				
7	2.2659	1.1052	0.8690	0.7020	0.5249	0.3259	0.1106			
8	2.2656	1.0956	0.8695	0.7303	0.5936	0.4409	0.2719	0.0919		
9	2.2649	1.0863	0.8639	0.7407	0.6306	0.5108	0.3770	0.2313	0.0780	
10	2.2641	1.0781	0.8561	0.7420	0.6493	0.5528	0.4454	0.3270	0.1998	0.0672

Simple LC Harmonic Oscillators

A simple, non-mathematical, discussion of ten single-stage LC feedback oscillators.

Dan Peters
Newport Beach, CA

When deciding upon a particular LC oscillator type, the designer may feel he has too many choices and ask, "Why do we need so many?" Most designers select a few favorites and stick with them, often willing to defend vigorously their particular, sometimes arbitrary, choices. However, even though you may only use a few, as a knowledgeable designer you should at least recognize the names of a number more.

In this discussion we limit ourselves to ten LC (as opposed to RC, crystal, etc.) oscillators of the feedback (as opposed to negative resistance, mode change, etc.) type and those that are further known as harmonic (as opposed to linear, square wave, etc.) oscillators and continuous (as opposed to relaxation, pulsed, etc.). In addition we limit ourselves to oscillators that can be implemented with simple one-stage amplifiers.

The LC feedback oscillator can be summarized in the simplified form of Figure 1.

The oscillators contain the following: an LC network which primarily determines the frequency of oscillation; a device capable of supplying power gain, to make up for the losses of the LC network and supply power to the load; and a feedback network to feed energy back to the

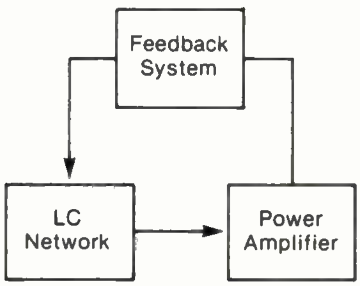

Figure 1. Generalized Feedback Oscillator.

LC network in the proper phase.

The Barkhausen criteria for oscillation states that the loop gain must equal one with a phase shift of 0°. If the gain is less than one, the oscillation will not sustain. With gain greater than one, the amplitude of oscillation increases until the amplifier limits or some other mechanism comes into play to reduce the gain to one. Knowing the gain versus amplitude characteristics of the various parts of an oscillator makes calculation of the amplitude of oscillation a fairly simple matter. The oscillator, when stable, must operate where the loop gain is one.

An early feedback oscillator, the **tuned-plate oscillator,** is shown in Figure 2. L and C constitute the LC network, the vacuum tube supplies the needed power gain, and the transformer action couples a portion of the plate energy back to the grid. The particular implementation of Figure 2 is grounded cathode, although it could also be grounded-grid or grounded-plate. It is series fed, but could be shunt fed. It also uses grid leak bias, but could use other types of bias.

The considerations of where and how to ground the circuit and how you want to feed the DC voltages are often very instrumental in the choice of oscillator type. We will ignore the particulars of DC supply and bias voltages. We will also leave out any reference to where the circuit is grounded. Our schematics will be simplified AC equivalent circuits and the circuit of Figure 2 reduces to that of Figure 3a. Figure 3b, the transistor equivalent, is known as the tuned-

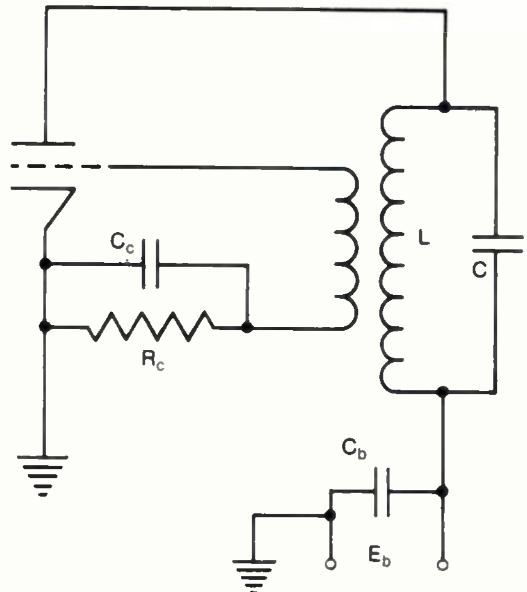

Figure 2. Tuned-Plate Oscillator.

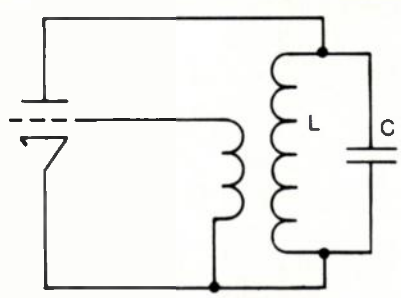

Figure 3a. Simplified Tuned-Plate Oscillators.

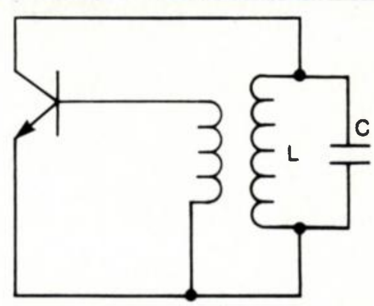

Figure 3b. Simplified Tuned-Collector Oscillator.

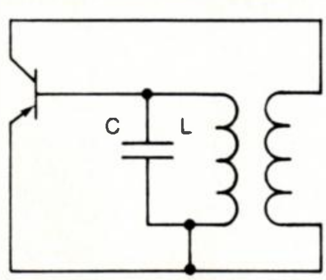

Figure 4a. Simplified tuned-base Oscillator.

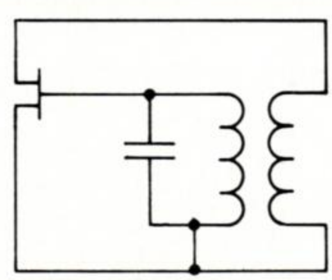

Figure 4b. Simplified tuned-gate Oscillator.

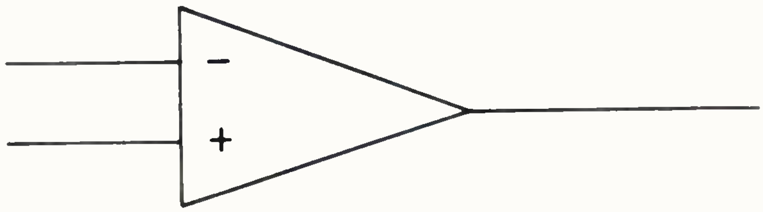

Figure 5. Generalized Amplifying Element.

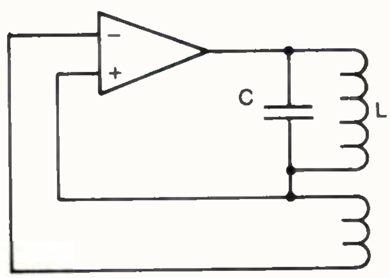

Figure 6a. Tuned-Collector Oscillator.

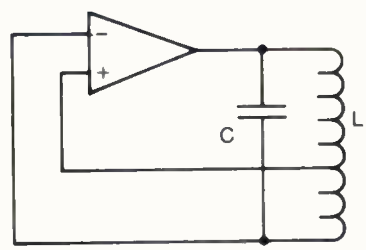

Figure 6b. Hartley Oscillator.

collector oscillator. Not shown is the FET version, known as the tuned-drain oscillator.

A very similar oscillator is the **tuned-grid,** or **tuned-base,** or **tuned-gate oscillator.** The tuned-base and tuned-gate versions are shown in Figures 4a and 4b. (Rather than show a vacuum tube, transistor, FET, or other amplifying element in our simplified schematics, we will use the symbol of Figure 5 to represent a generalized amplifying element.)

Tuning an LC oscillator over very wide frequency ranges is often accomplished with a variable capacitor and a number of different inductors. No matter how you elect to change the inductors, the fewer connections the better.

The **Hartley oscillator,** our next candidate, has three coinnections on the coil. If we re-draw the tuned-collector oscillator slightly, as in Figure 6a, it is not a big step to the *Hartley* oscillator, shown in Figure 6b. Although you can analyze the Hartley oscillator as a modification of the tuned-collector oscillator, it is interesting to note that the circuit will oscillate with no coupling between the two sections of the coil. However, you almost always see the circuit implemented with good coupling be-

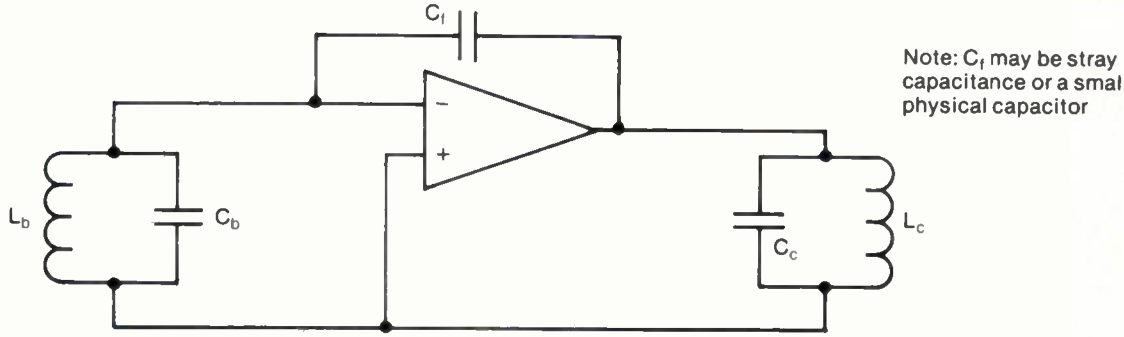

Figure 8. Tuned-base – Tuned-Collector Oscillator.

tween the two sections of the coil, because the overall efficiency and performance are considerably improved thereby. With coupling less than unity, the leakage reactance appears as an inductance in series with the emitter lead.

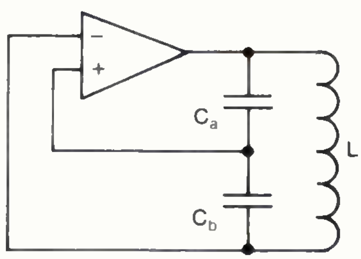

Figure 7. Colpitts Oscillator.

If you have no problem accepting the fact that the Hartley will oscillate with the two uncoupled coils, you will have no problem with our next oscillator. The **Colpitts oscillator,** shown in Figure 7, splits the capacitor rather than the inductor. An advantage of the Colpitts, over the preceeding oscillators, is that it has the simplest possible inductor. The price paid for this in a capacitor tunable oscillator is the requirement for a dual-gang tuning capacitor. If you attempt to vary frequency over more than a very narrow range by varying only one capacitor, the ratio between capacitors and, thus, the performance will change with tuning.

The tunable Colpitts often takes on a grounded-emitter configuration. This allows a common grounded rotor in the tuning capacitor, a convenient construction. A familiar application for the Colpitts is in Dip Oscillators, a common tool in RF labs.

The **tuned-base — tuned-collector oscillator,** shown in Figure 8, is probably built accidentally more often than intentionally. Some texts interpret this oscillator as a Hartley in which the mutual inductance has become zero.

The collector and base signals of a transistor are normally considered 180° out of phase, so at first look at signal fed from collector through C_f to the base may not seem to meet the Barkhausen requirement that the loop gain have a phase angle of 0°.

If you haven't already done so in the past, it is worth a few moments of doodling with some vectors and showing yourself that it is very easy to have the voltage fed back from the collector arrive at the base in phase with the signal present at the base. Allow for the phase shift across C_f and the phase shifts achieved by having the tuned circuits operating a little off their resonant frequencies. Do this one time and you won't be so surprised the next time a tuned amplifier oscillates on you.

Although the tuned-base — tuned-collector oscillator is seldom used in the form shown, it does find use where one of the LC circuits is replaced by some form of resonator, such as the **Miller crystal oscillator.**

The preceeding oscillators represent the prototype basic single-stage oscillators. Now let's look at some variations.

Before frequency synthesizers, the LC oscillator was *the* stable variable frequency source. As such, experimenters began to search for maximum frequency stability from LC oscillators. As they learned to improve the stability of the coils and capacitors used in their circuits, they realized that the amplifier itself was a major cause of instability. The vacuum tubes of the day drifted in gain, and the interelectrode capacitances varied.

One attempt at improving stability was to make the tuned circuit with very small coils and very large capacitors. In this manner, they hoped to make frequency variations due to the tube and stray capacitance variation minimal.

They soon realized that there were practical limits to this approach. Large C tanks are low impedance circuits, and there is a practical limit as to how low an impedance the amplifying element can work with. Also, large capacitance variable capacitors become impractically large physically. Because of these, and other problems, new approaches were sought.

One approach was to have the amplifying element not appear across the entire tuned circuit. The reasoning was, the less the LC circuit saw of the amplifier the less influence the variations in amplifier parameters would have on the LC circuit and the more stable the oscillator.

Three variations to this approach are the *Lampkin* oscillator, the *Meissner* oscillator, and the *Clapp-Gouriet* oscillator.

The **Lampkin oscillator,** designed in a 1939 Proceedings of the IRE, by G.F. Lampkin, is shown in Figure 9. You can recognize the Lampkin oscillator as a Hartley oscillator with

(Continued on page 61.)

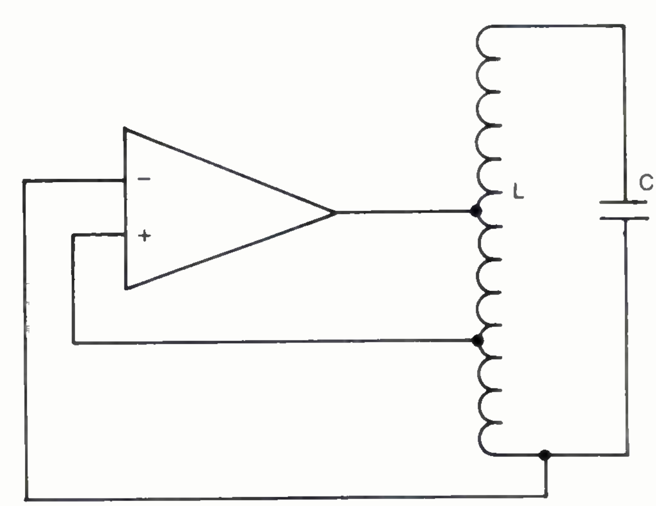

Figure 9. Lampkin Oscillator.

(Continued from page 58.)
the amplifier tapped down the coil. You can shift things around and slide things up and down without changing the basic principle.

A disadvantage of the Lampkin oscillator, associated with the large number of coil connections, is a propensity for the circuit to oscillate on some relatively high spurious frequency. Preventing spurious oscillations requires tight coupling between coil sections, not always easy at RF.

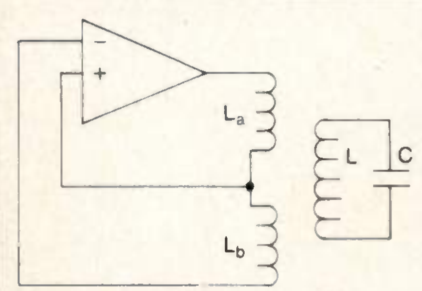

Figure 10. Meissner Oscillator.

Another attempt at isolating the LC circuit from the amplifier was by Meissner and is shown in Figure 10. Unfortunately, the **Meissner oscillator** displays even greater tendency toward spurious oscillation than the Lampkin, and the only application I found published for the Meissner was an interesting one where there was a need for a number of fixed frequencies to be achieved with plug-in modules. To avoid the well-known problems with plug-in contacts the design used a number of sealed plug-in modules. Each one contained a separate LC tank circuit and simply fit into a recess in the case where it magnetically coupled to L_a and L_b of Figure 10. L_a and L_b were fixed parts of the oscillator. Thus, no switched metallic connections were required.

The most popular of these original tap-em-down oscillators is known as the **Clapp-Gourlet,** or more commonly the *Clapp,* oscillator and was described by J.K. Clapp in a 1948 Proceedings of the IRE. The Clapp oscillator, a modified Colpitts, appears in Figure 11. You can tune the Clapp oscillator over reasonable frequency ranges by means of a single tuning capacitor (C_t). It is a popular choice for HF transmitter VFO's and HF receiver local oscillators.

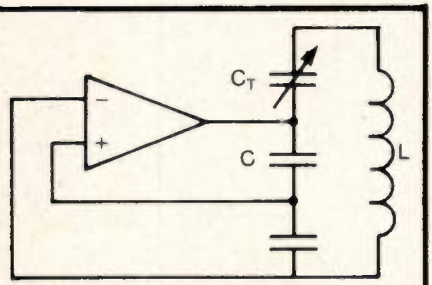

Figure 11. Clapp Oscillator.

An analysis of the frequency stability of the Colpitts oscillator and the Clapp oscillator will show that, when properly proportioned, the Colpitts oscillator stability can equal that of the Clapp oscillator.

However, the Clapp oscillator offers more flexibility. The inductance value may be chosen on the basis of convenience, compatibility with available tuning capacitors, etc. The impedance level presented to the amplifier may then be adjusted by C_c and C_b. In this way, the important engineering parameters are under good engineering control.

On the other hand, in the Colpitts oscillator the reactances required for stability are often impractically small, and an attempt to realize the calculated values of stability is frustrated by poor values of Q, impractically large variable capacitors, etc.

Tuning the Clapp oscillator by varying a single capacitor causes changes in output amplitude. A modification

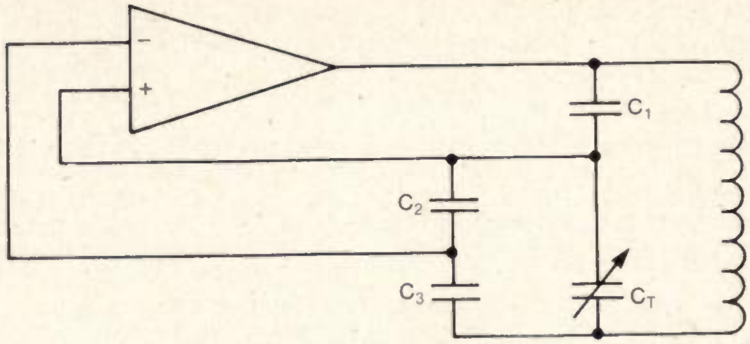

Figure 12. Vackar Oscillator.

of the Clapp oscillator which supplies relatively constant amplitude over wide frequency ranges is the **Vackar oscillator,** shown in Figure 12

Another variation of the Clapp oscillator is the **Seiler oscillator,** Figure 13. the Seiler can have slightly better stability than the Clapp or Vackar in practical circuits. Notice that if the Seiler is built in a grounded-emitter configuration, there is the advantage of having both one side of the tank coil and the tuning capacitor at ground potential. □

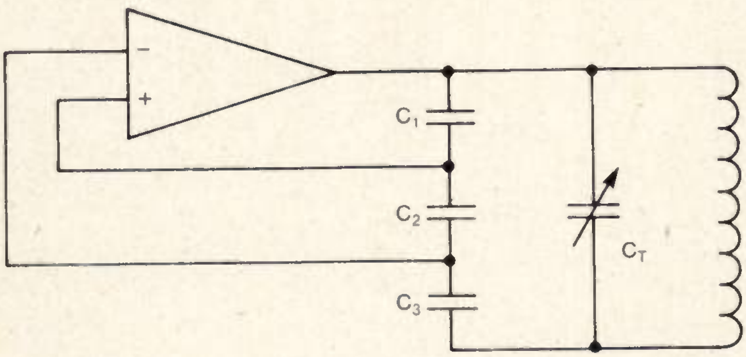

Figure 13. Seiler Oscillator.

News

Boulder, Colo., Aug. 18-20, 1981, "1981 IEEE International Symposium on Electromagnetic Compatibility." Contact: H.E. Taggart, Chairman, National Bureau of Standards, Boulder, CO 80303. (303) 497-3462.

Jerusalem, Nov. 9-12, 1981, "Isratech 81," Israel's high technology industry trade fair, held once every three years, will be held this year at the Binyanei Ha'ooma Convention Center in Jerusalem, Israel from Nov. 9-12, 1981.

More than 200 exhibitors of high technology products are expected to attract thousands of visitors from Israel and abroad. Showcased products will include but not be limited to the following categories: **Electronics:** Communications Equipment & Systems, Computer Hardware & Software, Industrial Controls & Equipment, Medical Electronics, Test, Measurement & Instrumentation Equipment.

"Isratech 81" visitors will have the opportunity to review the new products, to meet individually with manufacturers and distributors, to visit manufacturing facilities and to learn of the many advantages of trading with Israel, among which is duty-free trade for more than 2500 products with the EEC and the U.S. Contact: Joan Leavitt, Ruder & Finn, Suite 270, 1225 19th St. N.W., Washington, D.C. 20036. (202) 466-7800.

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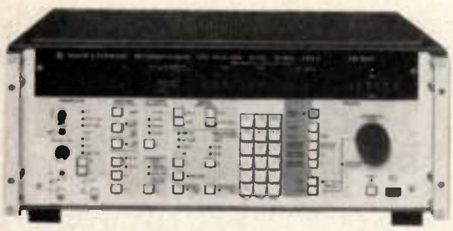

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Microwave Filters, Impedance Matching Networks, and Coupling Structures

G. Matthaei, L. Young & E.M.T. Jones

Ask any RF designer in the 1960's which book he regarded as the bible of RF/microwave circuit design and invariably he would mention "you know, that big thick one by Mattaei, Young and Jones." Any engineer who had this massive 1096 page volume was indeed fortunate and had the makings of an excellent library. The reason was obvious — 17 chapters of succinct and comprehensive design data for many disciplines. Consider the following examples: Ch 3 "Principles of the image method for filter design"; Ch 4 "Low-pass prototype filters obtained by network synthesis methods"; Ch 6 "Stepped-impedance transformers and filter prototypes"; Ch 12 "Band-stop filters"; Ch 14 "Directional, channel-separation filters and traveling-wave ring-resonators."

Available from Artech House, Inc., 610 Washington St., Dedham, MA 02026. 1096 pages, $50.00 cloth.

Ferromagnetic-Core Design & Application Handbook

M.F. "Doug" DeMaw

Ulrich Rohde, Ph.D., who needs no introduction, has submitted the following book review:

This handbook of ferromagnetic core design is an absolute must for any engineering library and is the best, if not the only, complete application book on this subject. The author also went through great pain and detail to inform the reader about the capabilities and applications of ferrite material and his impressive literature references indicate that the field was well searched.

The book has five chapters and five appendices dealing with the Basics of Magnetic Materials, Application of Rods, Bars and Slugs, Applying Toroidal Cores, Beads, Sleeves and Pot Cores, and Permanent Magnetic Data. The appendices provide various manufacturer supplied data on mechanical construction and electrical information.

The mathematical level in the book is sufficient for all practical applications and the necessary background and formulas are provided.

Traditionally, in American literature there is a large emphasis on toroids and toroidal cores, and it is good to see that pot cores are also covered.

It is customary to add a few remarks about possible improvements, and I would like to recommend the following to the excellent presentation:

1. Test results on intermodulation distortion on the various ferrite materials.
2. Information on the complex permeability.
3. Reference to some non-American papers that contain some other useful high-frequency applications.

Available from Prentice-Hall, Inc., Englewood Cliffs, NJ 07632. 256 pages, $19.95 cloth.

Handbook of Electronic Design And Analysis Procedures Using Programmable Calculators

Bruce K. Murdock

This book provides programs for programmable calculators enabling solutions to problems in network analysis, active and passive filter design, electromagnetic component design, high frequency amplifier design and engineering mathematics.

These programs are meant to serve at least two purposes. The first is to provide the working engineer or scientist a library of programs for his or her fully programmable calculator (e.g. HP-67/97 and TI-59) and secondly to provide by way of example programming tips and algorithms to enhance the usefulness of same. One way to gain programming proficiency is to study how others attack and program a particular problem.

Upon examining a "typical" program e.g. program 2-1 "Butterworth and Chebyshev Filter Order Calculation" this is what you'll find: 11 pages of "program description and equations used", Butterworth/Chebyshev nomographs, full page of *detailed* user instructions, four separate examples and two full pages of program listing replete with register, label, flag delineations.

Available from Van Nostrand Reinhold Co., 135 West 50th St., New York, N.Y. 10020. 525 pages, $29.50 cloth.

Radio Handbook

William I. Orr

This is the 21st edition of a handbook that has become a catchword among many thousands of electronic designers who also have in common the hobby of amateur radio. Do not think for a minute that there is anything amateurish about this 35 chapter, 1135 page volume. It is a comprehensive and clearly written text on many different areas of electronic theory and practice. It does not sacrifice accuracy for clearness. Listing just a few of the 35 chapters will give you a good appreciation of the author's intent: "DC Circuits," "AC, Impedance and Resonant Circuits," "Semiconductor Devices," "Vacuum Tube Principles," "SSB Transmission and Reception," "Communication Receiver Fundamentals," "Frequency Synthesis," "Amplitude Modulation and Audio Processing," "Radiation and Propagation, etc.

Available from Editors and Engineers, division of Howard W. Sams & Co. Inc., 4300 West 62 St., Indianapolis, IN 46268. 1135 pages, $21.50 cloth.

Microwave Circuit Design Using Programmable Calculators

J. Lamar Allen, Ph.D. and Max Medley

Network analysis programs covering the RF spectrum from VLF through millimetric waves are provided for the designer who has an HP67/97, HP-41C or TI-59 programmable handheld calculator. This book contains more than just program listings, it is tutorial by nature. It remands the designer to be efficient since he is using a limited program and data memory device (as only contrasted against large time-sharing computers). It accomplishes this by the following organizational divisions: "TI-59"; "HP-67/97"; "HP-41C Compatibility"; "TI-59 Program Listings"; and "HP-67/97 Program Listings." Closer examination of the "TI-59" division indicates the following sections: "Two-Port Analysis"; "Interconnections and Parameter Conversions"; "Lumped and Distributed Elements"; "Special Two-Ports"; "Mapping and Configuration Conversion"; "Ladder Network Analysis"; and a "Miscellaneous" Section.

Zeroing in on the "Mapping and Configuration Conversion" section indicates the following considered subject areas: "Two to Three Port Conversion," "Parameter Conversion," "Three to Two Port Conversion," "Three-Port Mapping and "Map-Source and Map-Load."

Available from Artech House, Inc., 610 Washington St., Dedham, MA 02026. 297 pages, $40.00 cloth.

Isolated BNC Dual-Line Protectors

Model C10 is designed to protect up to 10 data line pairs employing BNC connectors that connect to computers, modems, terminals, and other sensitive electronic equipment from the effects of transients caused by lightning, switching surges, heavy machinery, etc. The protector interfaces between the signal lines and the sensitive circuits to provide a

sophisticated blend of high speed (nanoseconds) voltage limiting and brute force protection. The signal line protector recovers automatically to standby in preparation for further protection. Clamping can be provided from 6 volts to 200 volts, depending on customer requirements.

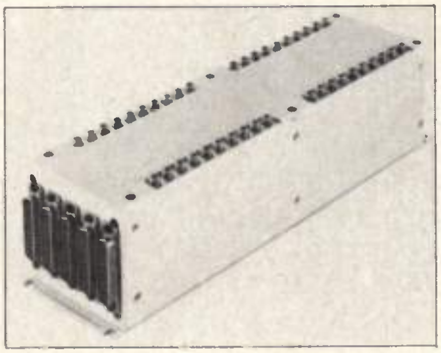
The dual line protector has a clamp voltage to ± 50 volts (in 5-volt steps), an energy handling capacity of 50 joules (min)/circuit, and a maximum frequency to 3 MHz.

Contact MCG, 160 Brook Avenue, Deer Park, NY 11729. INFO/CARD #140.

New 5 by 20 Switching Matrix

Matrix Systems Corporation has added model 6190 to its line of standard coaxial switching components. Model 6190 offers a 5 by 20 video switching matrix which is ideal for switching base band video and the HF bands.

The new model provides high isolation switching of video signals, and



may be controlled with 28 VDC or through a computer by using Matrix logic interface modules. The 6190 interfaces with SMA-type connectors; a mating connector is supplied as standard equipment. The 6190 is field repairable, requiring only a screwdriver and soldering iron. Reliable hermetically sealed reed relays are employed for switching speeds of one millisecond. Both dry and mercury wetted contacts are available.

Matrix Systems Corporation manufactures a wide line of modular coaxial audio switching equipment which may be computer or manually controlled. The company specializes in custom configurations.

Contact Matrix Systems Corp., 5177 N. Douglas Fir Road, Calabasas, CA 91302. INFO/CARD #139.

New Ultra-Broadband Voltage Controlled Oscillator

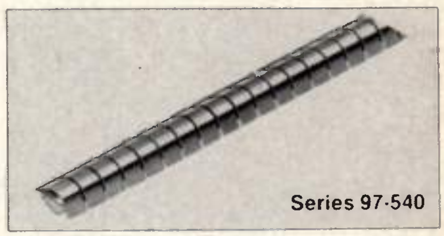
The Model V202 is a unique broadband oscillator that covers the entire 10 to 1,000 MHz frequency range in a single smooth tuning voltage controlled oscillator (VCO). This oscillator is a

miniature (2" x 2" x 1") precision self-contained assembly that operates from a single +24 volts, requires −24 to +24 volts to tune the entire band, and delivers +13 dBm across the band. This oscillator's 6 octaves frequency coverage compliments modern broadband components and fast ECL logic circuitry.

Contact Luff Research, P.O. Box 449, Jackson Heights, NY 11372. Circle INFO/CARD #138.

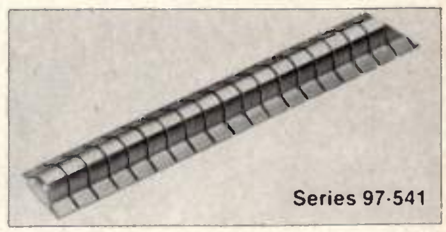
Extremely Narrow Electronic Shielding Strips

Two new electronic shielding strips in extremely narrow widths have been announced by Instrument Specialties Co., Inc. Precision-manufactured from



Series 97-540

beryllium copper, backed with self-adhesive, the new strips are only .28" wide in their free position, and .37" when compressed. Designed to shield cabinet doors, panels, and other enclosures for electronic equipment, the new strips are additions to the company's line of Sticky-Fingers® electronic gaskets. Each strip has its own self-adhesive backing, which pro-

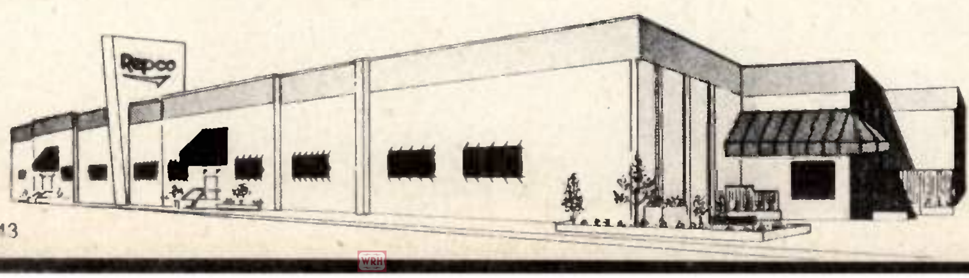


Series 97-541

vides a tight, instant bond without the need for holes, screws, soldering, or other fastening devices.

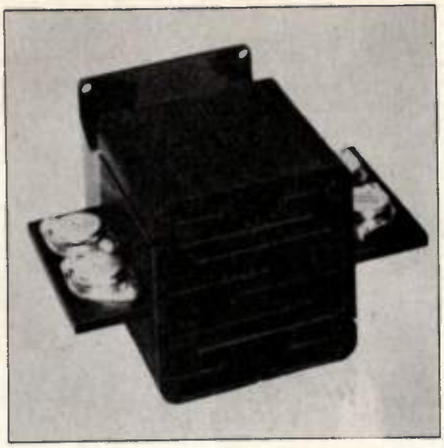
The new strips exhibit shielding effectiveness comparable to the rest of the line — which has measured greater than 46 dB at 14 kHz magnetic, and 108 dB at 10 GHz plane wave. They are unaffected in temperature ranges from −40°F to 250°F.

Contact Instrument Specialties Co., Inc., Delaware Water Gap, PA 18327. INFO/CARD #137.



Heat Sink Plus' 6031 Series

Heat Sinks Plus' 6031 Series forced air cooling package is a two piece modular constructed unit designed for mounting and cooling electronic components. The 6031 Series is constructed from Heat Sinks Plus' 6031 extrusion with a fin design that offers less air flow resistance resulting in maximum heat transfer and optimum thermal performance. The 6031 Series is specifically designed to be used with a mini boxer type fan (3.200" square) and the three inch quadrant shown will accept two TO-3 style devices for a total of four resulting in a maximum input with minimum temperature rise.



Contact Heat Sinks Plus, 28042 Del Rio Road, Temecula, CA 92390. Circle INFO/CARD #136.

Coaxial Cable Tap and Passive Fiber Optics Splitter/Combiner

Brochure 4484-0 from AMP Incorporated describes a new coaxial tap and passive fiber optic splitter that can be used together in a hybrid coaxial cable/fiber optic local data network that overcomes the limitations of pure coaxial cable and pure fiber optic data highway systems.

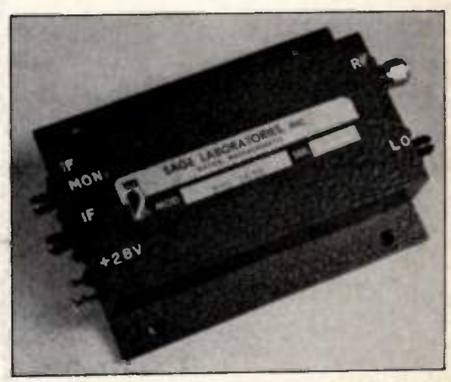
The coaxial tap can be installed at any location along a 5,000 meter cable without disrupting service and exhibits only 3pF loading. High impedance electro-optic transceiver circuitry can be housed in the tap body and driven by low voltage power carried on the coaxial cable.

The passive fiber optic splitter/combiner routes optical signals to, or from, seven remote units. Using this splitter/combiner, 19 coaxial taps can service up to 133 satellite locations, with each remote unit located up to one kilometer from the cable.

Contact AMP. Inc., 449 Eisenhower Blvd., Harrisburg, PA 17105. Circle INFO/CARD #134.

Dual Output Mixer Preamplifiers

Sage Laboratories introduces a new series of mixer preamplifiers that offer an overlapping coverage for signal frequencies from 600 MHz through 2.8 GHz.

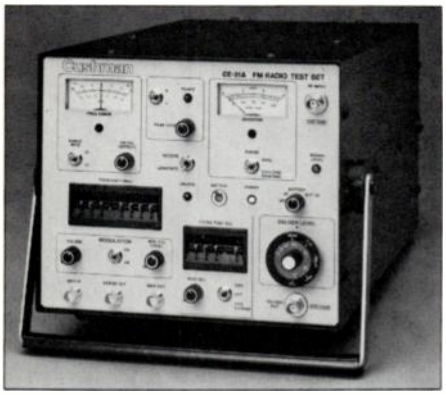


The signal frequency range for the FMA2638-1 is 0.6-1.25, the FMA2638-2 is 1.05-1.75, and the FMA2638-3 is 1.6-2.8 GHz. The units standard IF is 70 MHz and 1 dB IF bandwidth is 20 MHz. There are two isolated 50 ohm outputs at 70 MHz.

Contact Sage Laboratories, Inc., 3 Huron Drive, Natick, MA 01760. Circle INFO/CARD #133.

FM Radio Test Set

Cushman Electronics, Inc., has released one of the lightest weight, portable FM Radio Test Sets ever — the CE-31A FM Radio Test Set. At a featherweight 26 pounds, the CE-31A FM Radio Test Set contains in a single unit the test instruments needed to check the operation of an FM radio transmitter and/or receiver in THE VHF/UHF range of 25 MHz-999.9999 MHz. The RF Signal Generator section



can check receiver sensitivity from 0.1 microvolts to 10 millivolts. The CE-31A FM Measurement section can check a transmitter output frequency for accuracy and determine the output modulation deviation or phase angle. A demodulated signal output can be used to drive an oscilloscope to visually check modulation waveforms for distortion and accuracy.

Available options for the CE-31A are a Continuous Tone Coded Squelch System (CTCSS) sub-audible tone generator to open receiver squelch circuits, and the capability of using an internal battery and external +12 volts DC power source.

Contact Cushman Electronics, Inc., 2450 N. First Street, San Jose, CA 95131. INFO/CARD #132.

Technical Report On Termination Insensitive Mixers

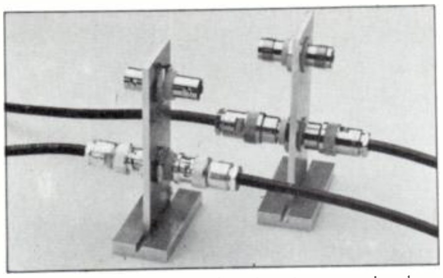
Anzac Division of Adams-Russell has prepared a 16-page report titled "Meet Tim" to describe how and why Termination Insensitive Mixers outperform other mixer types in real system use. The report features circuit design analysis, performance comparisons as well as actual test results, unit specifications and mechanical outlines. The report compares the TIM devices to other double balanced mixers and explains why, under system mismatch conditions, the Termination Insensitive Mixers outperform the others for Intermodulation, Conversion Loss Flatness and Spurious Signal Generation. Detailed specifications covering DC-8.0 GHz are provided.

Contact Anzac Division, Adams-Russell, 80 Cambridge Street, Burlington, MA 01803, for free copy. Circle INFO/CARD #131.

BNC, TNC Isolated Feed-Through Adapters

Two families of isolated feed-through adapters that eliminate ground loops and common mode current in coaxial transmission systems have been developed by Bunker Ramo Corporation's Amphenol North American Division. Designated the Amphenol• 31 Series, the isolated feed-through adapters

(U.S. Patent Pending) are available in BNC and TNC configurations to meet a broad range of applications requirements in analog and digital instrumentation equipment, television broadcast equipment, computers, telephone and security systems, aircraft and missile instrumentation, telemetry and general communications equipment. Reason for development of the BNC/TNC isolated feed-through adapters is the necessity of eliminating undesirable external noise from the outer coaxial cable conductor, introduced as a result of ground loops or common mode current created at



grounding points in the transmission system. The feed-through adapters eliminate ground loops by providing an uninterrupted path and continuous shielding between the signal source and the load.

Amphenol 31 Series BNC and TNC isolated feed-through adapters exhibit identical electrical and environmental characteristics. Both adapters are of the non-impedance type with voltage rating of 500 V peak. Center contact resistance is 1.5 milliohms, dielectric withstanding voltage is 1000 V RMS, and insulation resistance is 5000 megohms minimum.

Contact Amphenol North America Division Headquarters, Bunker Ramo Corporation, 2122 York Road, Oak Brook, IL 60521. INFO/CARD #128.

New Digital Inductance Substituter

IET Labs is proud to introduce a new user oriented Digital Inductance Substituter, which now rounds off its line of impedance substituters. It provides a fast and simple means of precisely setting any inductance over the very wide range of 1 mH to 9.999H.

Designated as the Model LS-400 L-box, the new Digital Inductance Substituter utilizes four side-by-side thumbwheel switches. The desired inductance is simply dialed in, and it is ready for use.

The LS-400 is error proof since the inductance is set and read directly as an unambiguous number on color coded switches. Unlike conventional decade boxes with rotary or slide switches, there is no need to examine or sum a whole group of separate numbers.

Contact IET Labs, Inc., 761 Old Country Road, Westbury, NY 11590. INFO/CARD #129.

Tubular and LC Filters

A complete new line of tubular and LC filters has been introduced by Wavetek Indiana. These precision filters can be optimized for each filter application by Wavetek's sophisticated new CAD (computer-aided-design) system. Wavetek's standard tubular

and LC filters cover the frequency range from 2 MHz to 10 GHz with typical 0.01 dB Tchebychev or Butterworth responses. Filters that operate outside the 2 MHz to 10 GHz range and filters with Gaussian, Bessel or Linear Phase responses are available.

Also available are folding, right angle, multiple, and other special filter configurations. Tubular filters may be specified in bandpass and lowpass designs; LC filters may be specified in bandpass, lowpass, bandstop and highpass designs. Tubular bandpass filters have 3 dB bandwidth ranges from 1 to 100 percent of center frequency. LC bandpass filters have 3 dB bandwidth ranges from 2 to 125 percent of center frequency. Standard filters are available with from 2 to 12 sections, 50 or 75 ohm impedances, 17 connector types, and average power ratings up to 200W.

Wavetek's design and quality control standards permit these filters to be specified to meet severe requirements for temperature, humidity, shock and vibration — including MIL-E-5400, MIL-STD-202, MIL-E-8189 and MIL-F-18327. Delivery time is 4 to 6 weeks.

Contact Wavetek Indiana, Inc., 5808 Churchman, P.O. Box 190, Beech Grove, IN 46107. INFO/CARD #135.

Synthesized SSB Radiotelephone

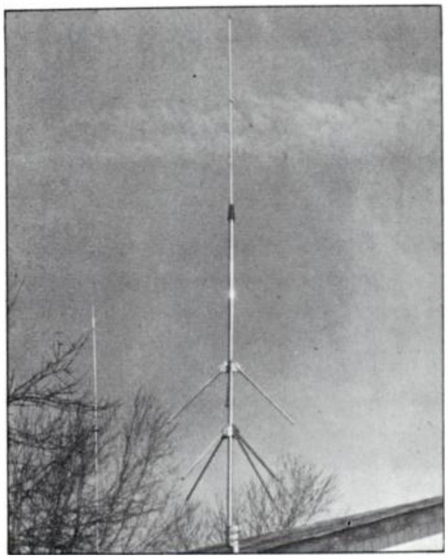
A new fully synthesized single sideband radiotelephone that features keyboard selection of any desired frequency between 1.6 MHz and 30 MHz as well as fast access to 192 pre-programmed ITU channels and 10 factory pre-programmed frequencies is now being marketed by Intech. For even greater convenience, an optional PROM circuit module allows storing of an additional 15 random frequencies for quick channel access. Designated the Intech Mariner 3600, this ruggedly built synthesized single sideband has a maximum power output of 150 watts making it an ideal radiotelephone for world-ranging marine vessels, vehicles traveling in remote areas and fixed base stations with worldwide communications responsibilities.

Keyboard entry of all selected frequencies or channels is microprocessor controlled. Each frequency is phased locked to a highly stable temperature compensated crystal oscillator that permits a precise 10 Hz resolution between individual channels. Selected frequencies and channels are displayed on a bright, easy-to-read digital readout display. An illumination control regulates brightness. A unique synchronous AM detector provides superior AM/DSB and AM/SSB reception while an audio derived squelch circuit substantially reduces operator fatigue.

Contact Intech, Inc., 282 Brokaw Road, Santa Clara, CA 95050. Circle INFO/CARD #126.

Double Zepp Antenna

Telex Communications, Inc. has an-



nounced that its Hy-Gain division will guarantee the new 2-meter V-2 amateur antenna to "equal or surpass the electrical performance of any competitive two stacked 5/8 wave antenna, regardless of gain claims, or your money back." If not satisfied the purchaser is required to return the antenna to the place of purchase within 30 days.

The antenna is a 2-meter extended double zepp vertical consisting of two stacked 5/8 waves decoupled inside the antenna for complete weatherproofing. The decoupling system allows no RF on the coax feedline. The V-2 is a complete antenna that is easy to assemble and will mount on any mast up to 2" (50.8 mm) in diameter.

Two sets of 1/4 wave radials and a centered feedpoint produce an excellent radiation pattern that is very close to the horizon with a minimum of power loss into the sky. Radiation pattern testing was achieved on a ground-reflection-range designed according to IEEE standard 149-1979 and the test results of the V-2 and various competitive products are available from Telex/Hy-Gain.

The V-2 is designed to operate from 138 MHz through 174 MHz and obtains a VSWR of less than 1.5:1 at resonance and has a 2:1 VSWR bandwidth of at least 7 MHz. The antenna's isolation from the supporting mast is 20 dB minimum.

Contact Telex Communications, Inc., 9600 Aldrich Ave. So., Minneapolis, MN 55420. INFO/CARD #127.

Catalog Featuring Over 20,000 Short-Term Rental Items

Genstar Rental Electronics, Inc., one of the oldest and largest national firms specializing in the short-term rental of electronic equipment, has announced the publication of a new Rental Catalog. The catalog lists more than $45-million worth of electronic instruments for rent throughout North America. The company, popularly known as Rental Electronics or REI, maintains inventory centers in major locations of the United States and Canada and has a separate used equipment sales facility.

The catalog itself is quite easy to use and understand, as all items include a description plus the 30-day rental price. Items are catalogued alphabetically by product type and are divided into 47 separate categories, which range from amplifiers to test chambers. There are analyzers, generators and meters of virtually every type, as well as sophisticated instruments such as desktop computers, microprocessor instrumentation, PROM programmers, oscilloscopes, magnetic tape recorders, and data terminals.

Table of Contents for the Rental Electronics catalog is on the front cover. The publication also includes an index, with items sorted according to manufacturer, and complete details on the terms and conditions of renting are also provided. Telephone numbers are given for all the REI North American facilities.

Contact Genstar Rental Electronics, Inc., 19527 Business Center Drive, Northridge, CA 91324, for free copies. INFO/CARD #130.

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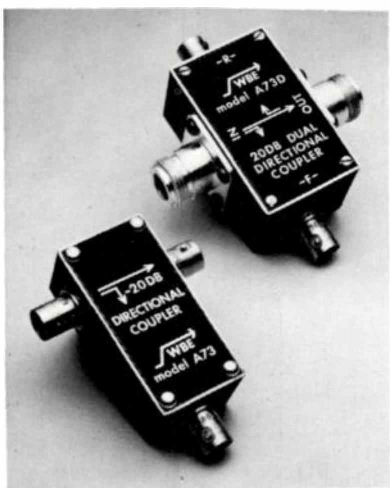

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A73-20GA				30	40			
A73-20GB				40	45			
A73-20P	1-100	single	50W cw (75 ohm limited to 10W cw)	35 dB min		.15	±.1	1.1:1 max
A73D-20P		dual		40 dB min typical		.3		
A73-20PX		single		45 dB min		.15		
A73D-20PX		dual				.3		
A73-20PA	10-200	single		35 dB min		.15		1.04:1 typical
A73D-20PA		dual		40 dB min typical		.3		
A73-20PAX		single		45 dB min		.15		
A73D-20PAX		dual				.3		

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