May 1980

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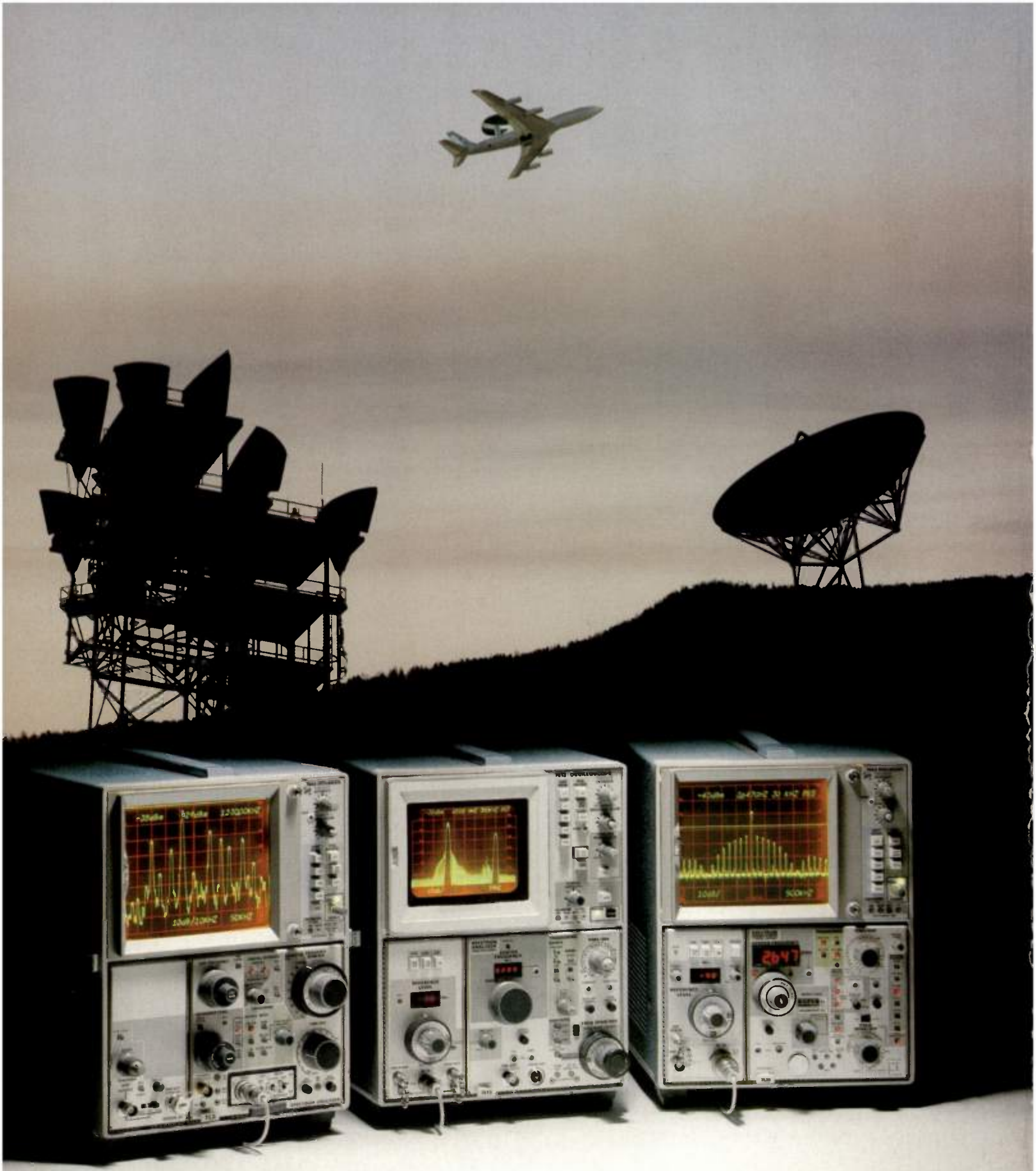

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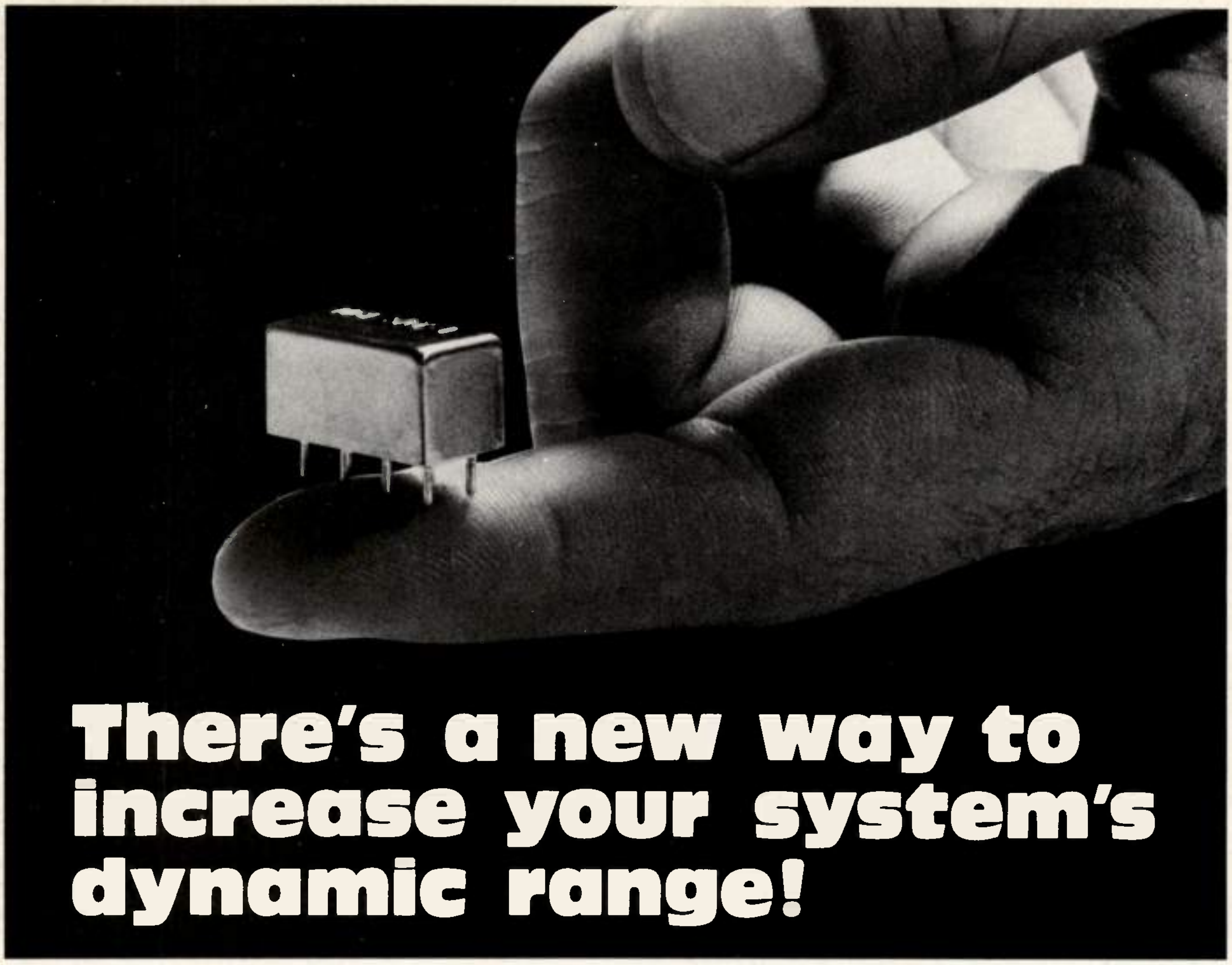

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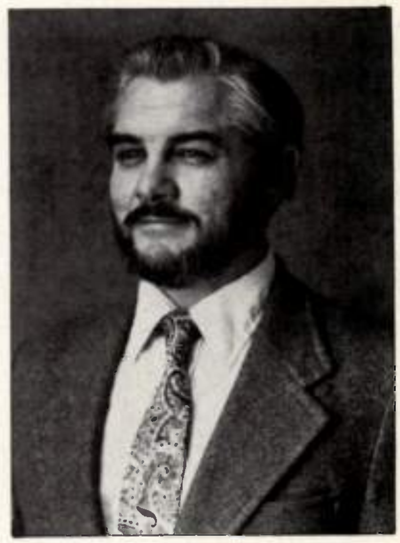
The Better Half

Every magazine needs a publisher and an editor. This editor you know from the February issue and subsequent editorials. It's time you met the man in the inset.

r.f. design is pleased to announce the appointment of Bill W. Childs as publisher. Bill comes to *r.f. design* with an impressive background in sales in the electronics industry. (This announcement is somewhat retroactive since Bill has been with *r.f. design* since February.)



Bill holds degrees in management, statistics, computer science and accounting. Bill has been in the electronics industry for the last 20 years. His experience has been in the aerospace and medical related aspects of this industry.

Early in Bill's career he became involved in computerized data reduction efforts from telemetry readouts on Polaris missiles. Later he participated in trajectory analysis for space shots and probability analysis on missile systems developments. Most recently, Bill has managed sales teams for selling major computer systems for hospitals and medicine. His major achievement in this area involved the sale of 15 million dollars worth of computer systems.

He managed a major systems implementation at the University of Colorado Health Sciences Center where a large scale patient registration, billing and information system was made operational in 1978. More recently, he directed the Western United States region for Technicon, a major producer of hospital information systems.

Bill resides high up in the hills of Evergreen, Colorado and through his extensive travel is by no means a stranger to your neck of the woods. Bill is on the West Coast this week and before long should be headed your way.

I am also very pleased to welcome Rob Coe, Doug DeMaw, Hank Keen and Andy Przedpelski to our Editorial Review Board. Their combined knowledge and experience is impressive and well received.

r.f. design tries to be responsive to the engineer, the engineering field and future trends. It is a consistently changing, on-going process that needs your participation. Let's perform like a Type 1 frequency-lock-loop. Input me with feedback, we'll integrate the results and be on the same frequency (or was that wavelength?).

Rich Rosen

Power Output Versus I_D

The broadband amplifier of Figure 2 performs over a 1 1/2 octave bandwidth with reasonably flat gain. Because of the usual idiosyncrasies inherent in the fabrication of matching transformers, the resulting performance of the amplifier showed a positive gain slope with increasing frequency.[7]

Fixing a quiescent drain current of 2.2 amperes (by setting the gate bias at 6.81 volts) and establishing an RF drive level of 8.5 watts, the measured performance across the 30-90 MHz band is tabulated in Table 1.

Table 1
Power Output & Drain Current vs. Frequency

Freq. (MHz)	I_D (A)	P_{out} (W)
30	4.99	84
40	5.13	89
50	5.25	95
60	5.43	100
70	5.58	102.5
80	5.83	105
90	6.00	107.5

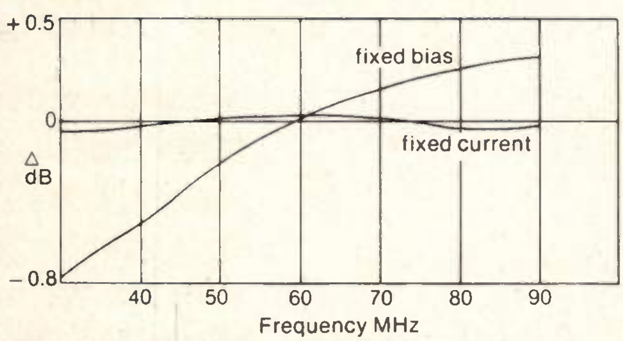


Figure 4. Fixed Bias & Fixed Current Power Output Vs. Frequency.

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7. ibid, Figure 2.

Note, from Table 1, that as the drain current increases with increasing frequency the power output follows suit. This current-power tracking is characteristic of VMOSFETs as a result of the linear transfer characteristics shown in Figure 1.

Since output power follows drain current we can presume that fixing the drain current will also stabilize the output power. Drain current was fixed at 5.46 amperes and the results are plotted in Figure 4 where they are compared against the power output when the drain current is allowed to float; that is, when V_{GS} is fixed.

AGC By I_D Control

Since power output can be stabilized by fixing the drain current which, in turn, can be regulated by gate bias, then a feedback mechanism working *entirely from the DC mains* can provide effective AGC action.

The simple AGC circuit shown in Figure 5 was built around the broadband amplifier of Figure 2. The circuit sensed drain current variations and adjusted V_{GS} as the frequency swept from 30 to 90 MHz. The resulting power output was leveled to within ± 0.1 dB as shown in Figure 6.

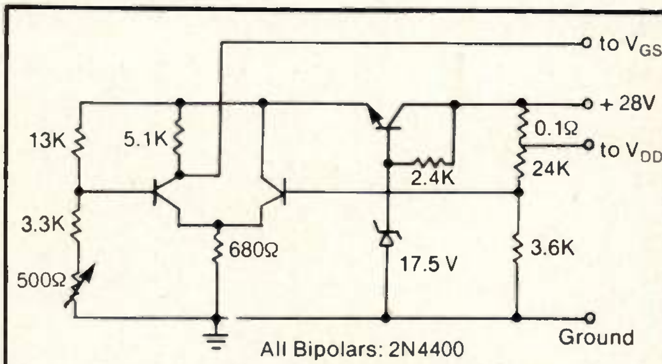

Figure 5. Simple AGC Circuit.

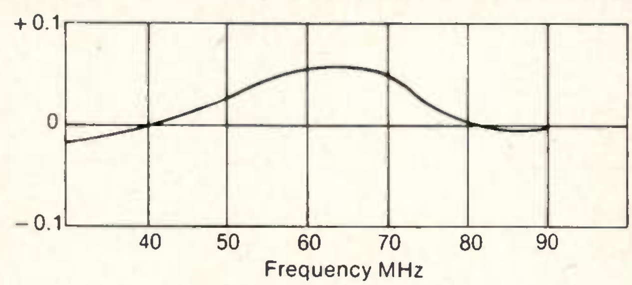

Figure 6. AGC-Controlled Power Output.

One precautionary note is that this AGC circuit senses the voltage developed across a 0.1 ohm resistor. In a laboratory environment one must guard that hook-up leads to and from the power supplies do not contribute to this voltage drop. Otherwise the AGC circuit will hopelessly try compensating for these losses.

Conclusions

AGC without RF sampling has been demonstrated to be effective when using VMOS in power amplifier design. Low-level gate modulation would be a natural outgrowth of this concept. □

RF Coaxial Cables — Choosing the Right One

A practical discussion that provides useful application information on the use, characteristics and benefits of the available coaxial transmission lines.

M.F. "Doug" DeMaw
Technical Department Manager, ARRL, Inc.
Newington, CT

"How long can I expect my RF transmission line to last? What effects will heat, soil acids and air contaminants have on it? What are the losses and will the cable handle the peak voltage it's subjected to?" These are common questions we engineers may ask, but the answers aren't usually found in reference books. So, let's examine some of the common problems in proper line selection and application, with the objective of making a correct choice for the specific task the cable will be called upon to perform.

Although coaxial cables are available for RF work with impedances from 25 to as great as 185 ohms, we'll key this discussion to the more popular lines which have characteristic impedances of 50 and 75 ohms. Emphasis will be placed on these cables in antenna systems, although the same rules apply for their use between pieces of RF equipment.

Impedance Considerations

The fundamental impedance of a coaxial line can be found from

$$Z_0 = \frac{L}{C} = \frac{138}{\sqrt{E}} \log \frac{D}{d} \text{ ohms} \qquad (1)$$

where D is the inner diameter of the outer conductor in inches and d is the outer diameter of the inner conductor in inches. To find C we can apply the equation

$$C = \frac{7.36E}{\log D/d} \text{ pF per foot} \qquad (2)$$

where E is the dielectric constant of the cable insulation between the conductors. Factor L is obtained from

$$L = 0.140 \log \frac{D}{d} \quad \mu\text{H per foot.} \qquad (3)$$

These equations can be applied if we are fabricating our own transmission lines from copper tubing.

It is important to realize that the actual terminal impedance of a specified line at a particular frequency may differ substantially from the design impedance (surge impedance). This can be affected by the voltage standing-wave ratio (VSWR) since the VSWR of a selected length of cable indicates the difference between the real input impedance of the line and the average characteristic impedance. This accounts for the variations seen with a VSWR indicator that is inserted in a line which has a VSWR condition (other than 1.0:1). If there is no appreciable VSWR present, the cable should indicate "flatness" at any point where the instrument is inserted.

VSWR uniformity is another significant impedance consideration. The actual VSWR of a cable section is the aggregate of random and periodic reflections inside the line. Connectors, line terminations and the cable itself contribute to the end VSWR condition. In applications where the VSWR amount is critical, such "bumps" in the line can be reduced or eliminated by purchasing the complete cable harness with connectors from a manufacturer who verifies its performance by sweeping the line assembly. The radial bends in some types of cable can introduce significant bumps in the line impedance (especially if aluminum-jacketed Hardline is employed and allowed to kink or crimp even slightly). Performance, in this case, deteriorates with increasing frequency.

Still another parameter we must consider is the *impedance* stability of the line we intend to use. It can be seen from (1) that the capacitance of the line figures significantly in the Z_0 determination of the cable. Some of the softer dielectric materials permit changes in cable capacitance when the cable is stressed or subjected to temperature extremes. This becomes especially critical at higher operating frequencies, above, for example, 900 MHz when short lengths of line are used. Similarly, if the dielectric of the cable is allowed to move, within the cable connectors, there is a change in capacitance, and hence a line bump. In the interest of minimizing changes in cable capacitance with temperature or stress, semiflexible coaxial lines are recommended, such as Times Wire and Cable Alumispline. Alumifoam or corrugated Alumifoam T[1]. The general structure of these lines is illustrated in Figure 1. The smaller foam-dielectric flexible lines are perhaps the worst offenders for the foregoing condition. However, most quality coaxial cables will maintain their specified impedances and capacitances within 2 percent over the manufacturer's stated operating temperature range when long sections of cable are used.

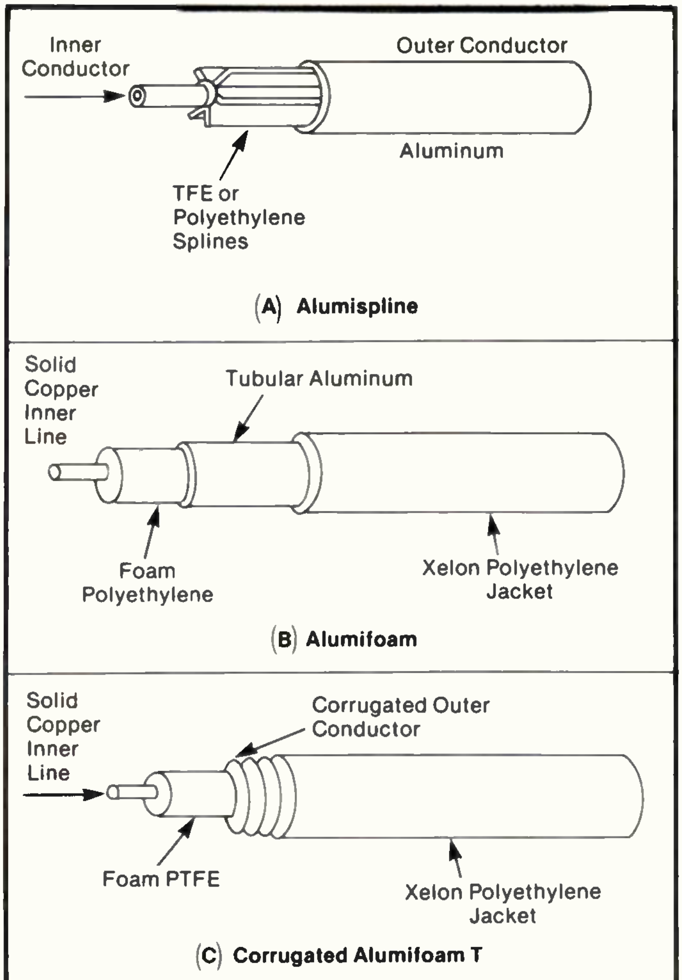

Figure 1.

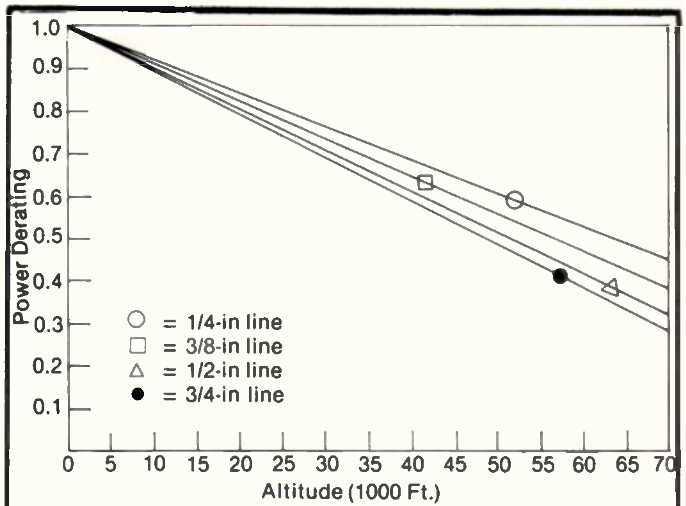

Figure 2.

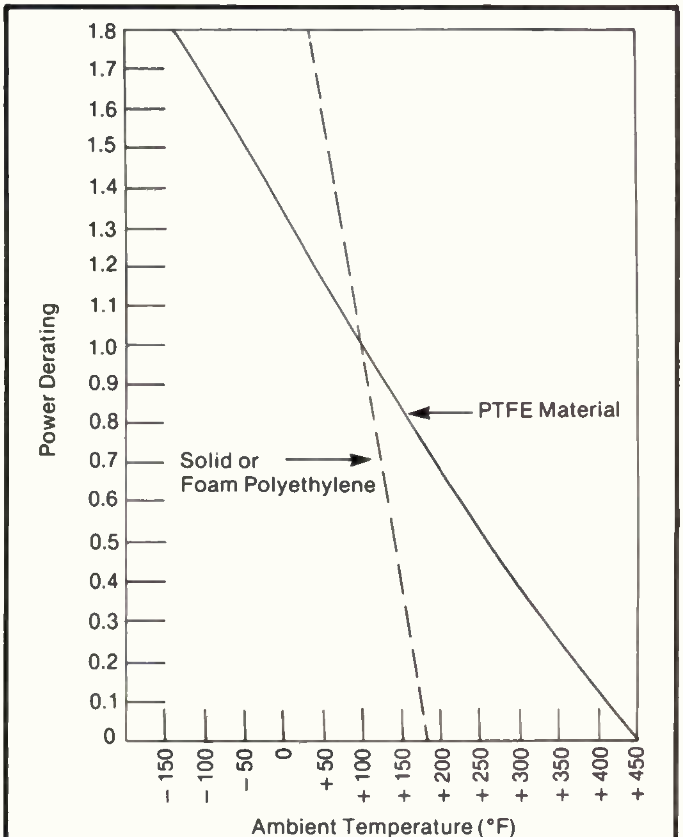

Figure 3.

What About Power Rating?

The major factors which affect the power rating of a specified coaxial cable are heat, altitude and VSWR. VSWR causes hot spots along the line. High ambient temperatures and high altitude both lower the power rating of the cable by preventing internal heat from escaping from the line at a normal (required) rate. Therefore, all of the conditions must be considered when selecting a cable for a particular environment.

The correct procedure for choosing a proper cable is to first determine *average* input power to the line at the highest anticipated operating frequency. Next, calculate the *effective input power by*

$$P_{in(eff.)} = \frac{P_{av} \times \text{VSWR Correction}}{\text{Temp. Correction} \times \text{Alt. Correction}} \qquad (4)$$

Corrections for altitude and temperature are provided in Figures 2 and 3. The VSWR correction factor can be found from

$$\text{VSWR corr. factor} = 1/2\left[VSWR + \frac{1}{VSWR}\right] + \qquad (5)$$

$$\frac{1}{2K}\left[VSWR - \frac{1}{VSWR}\right]$$

Acknowledgements

The author wishes to express his gratitude to Times Wire & Cable for the product samples with which long-term testing was performed by the writer in various environments. Appreciation is expressed also to Times Wire & Cable President Larry De George for his past counsel, along with that of the Times Wire Engineering staff.

Additional thanks are in order to Vice President Ed Rinehart of Decibel Products Corp. who generously provided cable samples and data sheets for their impregnated 50-ohm line.

where k can be found in Figure 4.

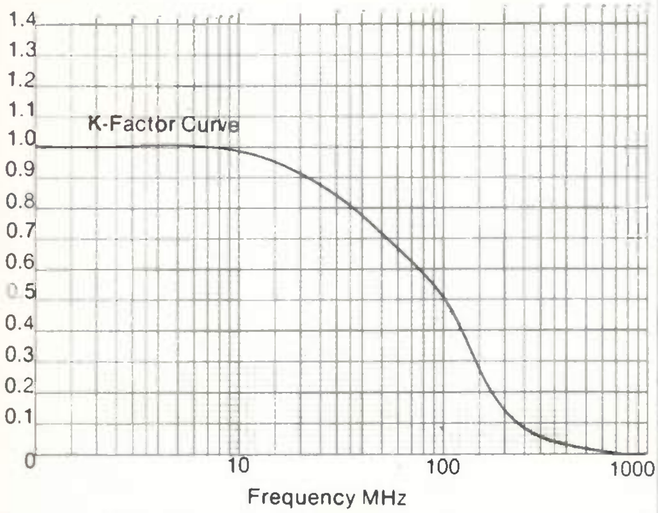

Figure 4.

Operating Voltage

Our concern when choosing a cable for a specified application must also take into account the AC operating voltage on the line. The proper approach is to calculate the operating RMS or peak voltage the system produces. Once this is known we can find the effective E_{in} by multiplying the actual E_{in} by the square root of the VSWR. A cable is then selected which has an ample safety factor for the system requirements.

The maximum operating voltage should always be less than the rated corona level for the cable (extinction voltage) since a continuous corona condition leads to dielectric breakdown, noise in the line and subsequent permanent damage to the dielectric material and the conductors.

The dielectric strength rating of the cable differs considerably from the corona level. It is a test voltage which the manufacturer applies for 60 seconds when the cable is manufactured. The DC voltage rating of coaxial lines is approximately three times the maximum permissible RMS voltage level. Most manufacturers provide information concerning the maximum safe AC voltage levels for their cables.

Velocity Factor

Each type of coaxial cable has a specific velocity factor which is dependent upon the dielectric material used between the inner and outer conductors. The relationship is defined as

$$V_f = \frac{100}{E} \qquad (6)$$

where V_f is the velocity factor as a percentage of the speed of light and E is the cable dielectric constant. The time-delay characteristic of a given cable can be determined from

$$T = 1.016\sqrt{E}\ \text{ns per foot} \qquad (7)$$

where T is the time delay in nanoseconds and E is the dielectric constant. Table 1 lists V_f and T for the various popular cables.

Table 1

Dielectric	V_f	T
Solid Poly	65.9%	1.54
Foam Poly	80.0%	1.27
Air-Space Poly	84-88%	1.15-1.21
Solid Teflon	69.4%	1.46
Air-Space Teflon	85-90%	1.13-1.20

It can be seen from the above that the electrical line length will be substantially different than the free-space resonant length. E.g., a one-half wavelength section of foam polyethylene-dielectric cable at 90 MHz will be

$$L(feet) = \frac{468}{f(MHz)} \times 0.80 = 4.16\ feet \qquad (8)$$

Similarly, the line length when using solid polyethylene dielectric will be 3.426 feet. The composite length of any line which uses cable connectors at the ends must include the portion of the connectors above the ground plane of the equipment, as seen in Figure 5. The coax connectors and barrel connectors must figure into the overall line length of a run of cable when joining one or more sections of cable (if electrical line length is a vital factor).

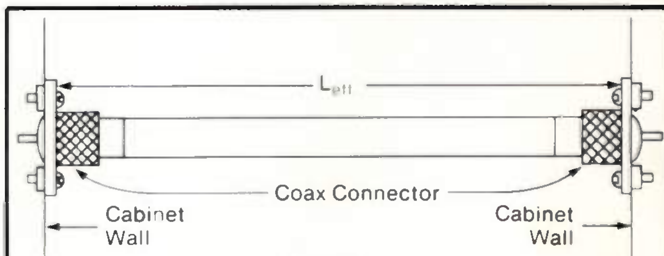

Figure 5.

Attenuation Characteristics

Perhaps one of the most important parameters worth considering when selecting a suitable cable for an application is the cable loss or attenuation. When the cable properties are known, the attenuation (α) is obtained from

$$\alpha = \frac{0.435}{Z_o(D)}\left[\frac{D}{d}(K1 + K2)\right]\sqrt{f} + 2.78\sqrt{E}(PF)(f) \qquad (9)$$

where Z_o is the line impedance, D is the inner diameter of the outer conductor, d is the outer diameter of the inner conductor and f is the operating frequency in MHz. K1 is the manufacturer's strand factor and K2 is the braid factor. PF is the power factor. This equation is based on dB loss per 100 feet of cable. It can be seen from (9) that the losses increase as the operating frequency is elevated. Therefore, it is important in terms of efficiency that we limit the line length (especially at VHF and higher) to only that amount which is necessary. Similarly, the quality of the line should be the best that the economics of the project will allow when long runs of cable are needed. The smaller the cable diameter the higher the attenuation, i.e., 100 feet of RG-58A at 100 MHz shows a loss of 4.9 dB as opposed to 0.85 dB with RG-231 aluminum-jacketed Hardline. The paradox connected with the use of long runs of small-diameter, lossy cable at VHF and higher is the illusion that a good

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match to the load exists, even though the VSWR may be relatively high (at the load), assuming of course that the VSWR measurement is made at the equipment end of the line.

The attenuation traits of coaxial cables often becomes more pronounced with *age*. This results from contamination of the inner dielectric material from an environment of moisture and high humidity, plus chemical action brought on by the outer jacket of the cable (plasticizer chemicals). This effect becomes of paramount importance at 1000 MHz and higher. Silver-plated copper shield braids fare considerably better in terms of longevity degradation than do cables which contain bare copper or tinned braid material. Therefore, if economics permit, it is best to buy the higher grade of cable when long periods of use are anticipated.

It is worth noting that foam polyethylene transmission lines with braided copper outer conductors exhibit roughly 15 percent less attenuation than cables of the same size and impedance which contain solid polyethylene dielectric. The "fly in the hand cream" is that the foam-dielectric lines become much more lossy than solid-dielectric ones when subjected to high levels of moisture. Hence, in areas with normally high humidity it is *inadvisable* to employ foam-dielectric line above 150 MHz. Similarly, try not to use cables which have PVC (type 1) outer insulation, as the chemical action with time contaminates the line and increases the attenuation. The most positive cure-all for the foregoing problems is to utilize the semi-flexible Hardline cables which contain foam dielectric. The solid-sheath aluminum outer conductors offer excellent protection against moisture and contamination. A suggested brand of this type of cable is Times Wire Alumifoam (see Figure 1). A less-expensive alternative is the use of flexible RG-8/U or RG-9/U style line which has been heavily impregnated with weather sealant. Decibel Products, Inc. manufactures an excellent transmission line of this type (VB-8).[2] It uses a solid center conductor as opposed to the normal stranded kind. This permits forcing the moisture-proofing agent between the center conductor and the dielectric material. Since the conductor is of solid copper, moisture and other contaminants can't seep into the cable via the center conductor. In a like manner, the impregnant is forced into the shield braid (bare copper) to form a moisture and chemical barrier. This type of line is suitable for being buried in the ground, as it is virtually impervious to soil acids and moisture. Furthermore, if rodents chew into the line (which is a common problem in some regions), the line is for all practical purposes, self-healing.

What About Shielding Integrity?

We're all familiar with the truism that "the structure of the outer conductor of a coaxial cable determines its shielding efficiency." But what is the effective shielding afforded by the various styles of braid? Well, single-copper-braid line provides 85 to 90 percent coverage of the line, irrespective of its being bare copper, tinned copper or silver-plated copper. Next up the scale comes double braid (two layers of braid with no insulation between them), which is substantially better than single-braid line. Next is triaxial braid (2 shield braids with insulation in between). There is also a "strip-braid" style of line. It is a shield braid which employs flat copper strips instead of small round wires. It provides some 90 percent of coverage.

The best line for shielding integrity is solid-sheath cable of the type illustrated in Figure 1. A one foot length of solid aluminum-jacketed line has a radiated power (referenced to desired signal power) which is *−700* dB at 100 MHz. Single-braid flexible line, by comparison at 100 MHz, has a radiation level which is only 48 dB below the desired signal level. Double-braid line, on the other hand, falls in between the two examples at − 98 dB for operation at 100 MHz.

Despite the integrity of the cable shielding it is always wise to isolate cables from one another as much as possible when they are routed up a tower or laid in a conduit. The fundamental rule for isolation of two or more cables is, "the total isolation (cross talk) between two coax cable runs is the sum of the isolation factors of the two outer conductors of the adjacent cables, and the coupling factor between the runs."[1]

Environmental Damage

The main causes of failure in coaxial cables, as a function of time, are the *moisture* we discussed earlier, *sunlight*, corrosive *salt-water vapor* and *galvanic action*. High-molecular-weight polyethylene dielectric is recommended for use in areas of high sun intensity and duration. PVC outer jackets should be avoided in preference to polyethylene jackets, as the life span of PVC in a high ultra-violet environment is approximately half that of polyethylene.

For installations near salt water it is best to use impregnated lines of the type discussed earlier. Times Wire & Cable *Imperveon* line is also excellent for preventing corrosion from salt air. Impregnated cable is recommended also in industrial areas where the air is saturated with pollutants.

As discussed earlier, heat has a pronounced effect on coaxial cables. The various insulating materials have different temperature profiles. Therefore, it is prudent to take the temperature factor into account when selecting a line for your application. Table 2 lists the typical operating-temperature boundaries for the most common dielectrics.

Table 2

Inner Dielectric Material	°C Range
Polytetrafluorethylene (PTFE)	− 250°C to + 250°C
Polyethylene	− 65°C to + 80°C
Foamed Polyethylene	− 65°C to + 80°C
Foamed or Solid Ethylene Propylene	− 40°C to + 105°C
Jackets	
Fluorinated Ethylene Propylene (FEP)	− 70°C to + 200°C
Polyvinylchloride (PVC)	− 50°C to + 105°C (depending on compound used)
Polyurethane	− 100°C to + 125°C (depending on compound used)
Nylon	− 60°C to + 120°C
Ethylene Propylene	− 40°C to + 105°C
High Molecular Weight Polyethylene	− 55°C to + 85°C
Silicone Rubber	− 70°C to + 200°C
Silicone Impregnated Fiberglass	− 70°C to + 250°C
High Temperature Nylon Fiber	− 100°C to + 250°C

References

1. *RF Transmission Line Catalog and Handbook*, Times Wire & Cable, 358 Hall Ave., Wallingford, CT 06492.
2. Decibel Products, Inc., 3184 Quebec St., Dallas, TX 75247. See VB-8 product review in *QST* Magazine for Nov. 1978, p. 38.

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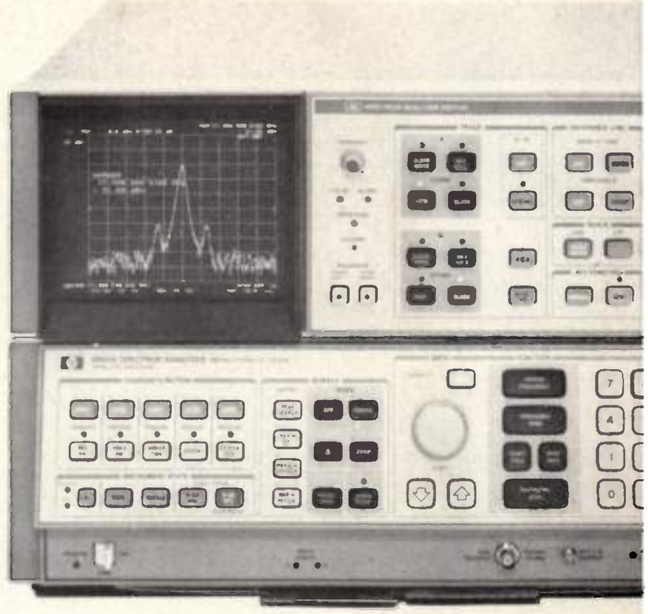
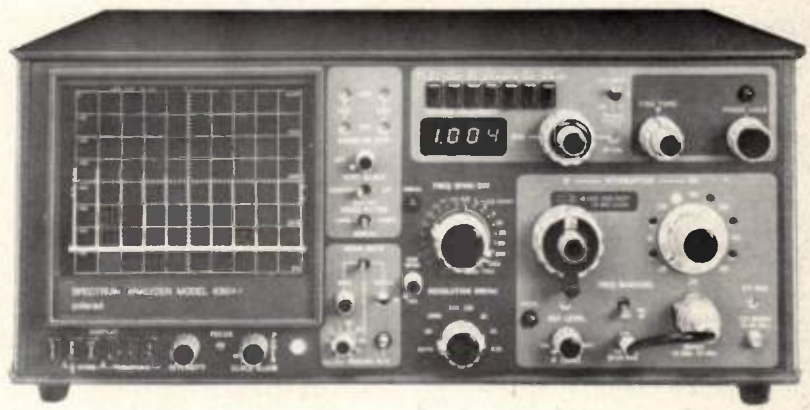
Spectrum Analyzers 1969-1979 A Technology Update





Advances in performance, packaging and ease of sweeping spectrum analyzers are presented. Future developments are indicated.

Len Garrett
Morris Engelson
Tektronix
Beaverton, Ore.

Spectrum analyzers available today are more accurate, cover a wider frequency range, and are easier to use than previous instruments. A look at the performance of modern units gives one an appreciation of these benefits (Figure 1). Looking also at current trends in packaging is equally enlightening. The packaging art has been taxed to the fullest today with some spectrum analyzers meeting the rugged MIL T–28800B specifications, plus being designed within a specified minimum size envelope. Finally, there has been a significant increase in ease of use. The number of interacting controls is being reduced drastically, making it easier to get desired measurement results. Prices are more realistic also. The key to value is to provide all the performance users need, without requiring them to pay for performance not needed. An example of this[1] offers as options a preselector, digital storage, programmability, frequency stabilization resolution, and provision for use of external mixers.

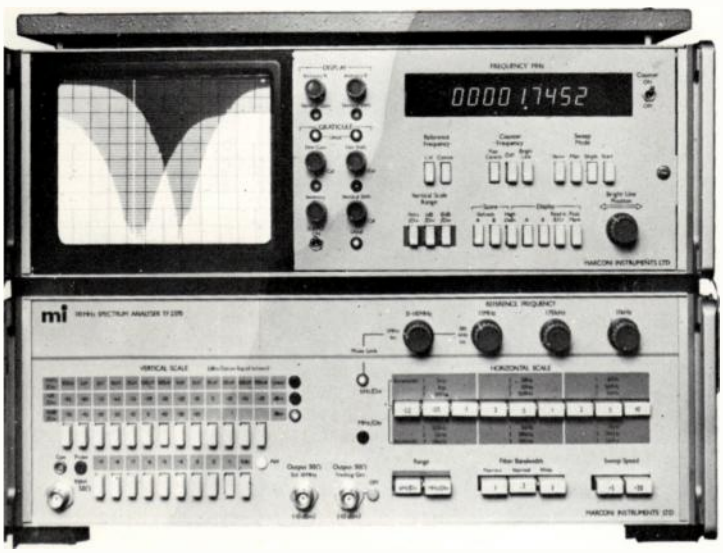
Significant advances in spectrum analyzers today are occurring in all areas: performance, packaging, environmental capability and ease of use. This is due primarily to the increased use of microprocessors, which are having a manifold impact in areas such as absolute and relative amplitude measurements, positive phase-locked loop control for never-break-lock operation, and very wide local oscillator ranges (Figure 2).

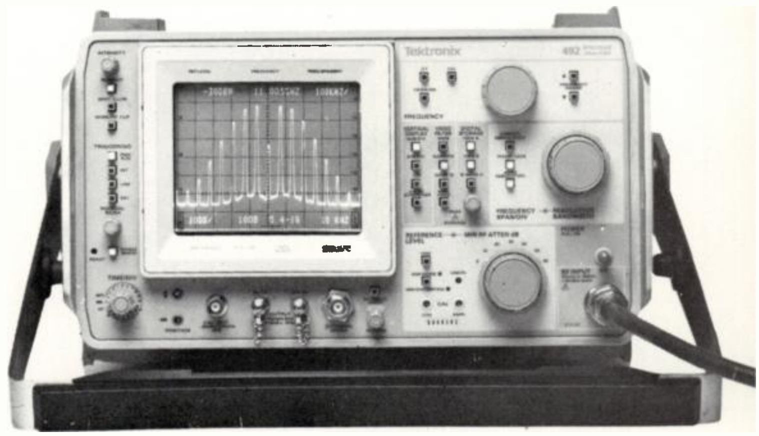
Increased versatility provided by the microprocessor enables spectrum analyzers to be used in an ever-widening range of applications:

- Oscillator spectral purity measurements
- Millimeter wave R&D
- Spurious response measurements
- Signal source stability measurements
- Signal source phase noise measurement
- Occupied bandwidth
- Surveillance
- RFI/EMI measurements
- Distortion measurements
- Pulsed RF source evaluation

Better Performance

Important performance parameters now improved for





spectrum analyzers are frequency range, resolution, stability, dynamic range, sensitivity, and amplitude accuracy. Newer models also include increased automation and programmability.

Frequency Range

The top frequency setting on commercially available spectrum analyzers has been 60 GHz. Options available today extend this upper limit to 220 GHz, while still maintaining a lowest characterized frequency of 50 kHz.

Two new technological factors contribute to this high-frequency-end breakthrough: broadband, full-amplitude-calibrated waveguide mixers operating to 60 GHz, and a 2 to 6-GHz local oscillator. Use of a 2 to 6-GHz oscillator rather than the traditional octave-range oscillator provides higher operating frequencies at lower harmonic conversion numbers.

Spectrum analyzers with full-frequency span and input frequency calibration features can take advantage of commercially available mixers, extending millimeter wave spectrum analysis to 220 GHz (See Figure 3).

Resolution

Resolution bandwidth available today ranges from 10 Hz to 3 MHz, with a fundamental conversion sensitivity of −123 dBm at 100 Hz. Combined with less than 50 Hz incidental FM and a wide-operating-range local oscillator, a 100 Hz resolution bandwidth setting is usable over a typical full coaxial input frequency range of 21 GHz. Respectable measurements at a 100 Hz resolution setting are possible even at 50 GHz. In addition, sideband noise is sufficiently low to permit easy observation of 70 dB down signals only 3 kHz removed from the carrier in a fundamental mixing mode.

Stability

Automatic instrument stabilization is assured by the microprocessor. Being able to stay phase-locked no matter what external temperature change or other adverse frequency destabilizing conditions are experienced is a real benefit. This performance is provided by the microprocessor's monitoring of various internal spectrum analyzer functions. Whenever frequency drift threatens to disturb phase lock, the microprocessor readjusts the local oscillator tuning voltage to maintain balance. The advantage of this system is that continuous frequency restabilization by the user is not necessary. Ease of measurement is greatly improved.

Amplitude Accuracy

Differential amplitude functions available allow amplitude comparison in 0.25 dB steps. This variation with the I-F substitution technique provides 200 one-quarter dB steps, allowing precision amplitude comparison over a 50 dB range. Here again, the result is improved ease of use and accuracy whenever relative amplitudes need to be determined, such as with AM, FM and pulse signals (See Figure 4).

Programmability

Control settings and signal display information can be

SPECTRUM ANALYZERS KEY FEATURES & SPECIFICATIONS

	TEK492P	H.P.8566A	EATON 757
Frequency Range	50 kHz to 220 GHz	100 Hz to 22 GHz	10 kHz to 22 GHz (3A)
Center Frequency Accuracy	±(.2% of center freq. + 20% of span./div.)	±[2% of span. width + (ref. x C.F.) + 10 Hz]	0.2% of center freq. (3B)
Resolution Bandwidth	(6 dB BW) 100 Hz to 1 MHz (1A)	(3 dB BW) 10 Hz to 3 MHz	(3 dB BW) 100 Hz to 1 MHz
Sensitivity Average noise level in 1 kHz BW @ 18 GHz	−90 dBm	−100 dBm	−90 dBm
Stability Fundamental Mixing	50 Hz P-P	0.2 Hz P-P (2A)	5.0 Hz P-P
Dynamic Range On Screen	80 dB	90 dB	100 dB
Drift After 1 Hr. Warm-Up Fundamental Mixing	±25 kHz/hour	±10 Hz/minute of sweep time (2B)	±3 kHz/10 minutes
Noise Sidebands 1 kHz res. BW & 10 Hz Video BW	70 dBc @ 30 kHz offset	75 dBc @ 30 kHz offset	70 dBc @ 30 kHz offset
Amplitude Measurement Accuracy typical	±4.3 dB @ 18 GHz	±3.2 dB @ 18 GHz	±3.0 dB @ 18 GHz (3C)
Automatic Measurement Capability	IEEE-488 Compatibility Full Program Control	IEEE-488 Compatibility Full Program Control	Data Out Only Freq. Programmable
CRT Display & Readout	Digital (1B)	Digital	Digital
Size (Height, width, depth)	6.9" x 12.87" x 19.65"	11" x 16-3/4" x 23.5"	8-3/4" x 16-3/4" x 21-7/8"
Weight	44 lbs.	112 lbs.	65 lbs.
Volume FT³	1.12	2.38	1.55
Environmental	MIL T 28800B Type III, Class 3, Style C	Mil Std. 461A EMI CE03, RE02	Not Specified
Power Consumption	215 watts	650 watts	220 watts
Price	$31.25K (1C)	$49.5K	$23.1K
Contributor of Information	Len Garrett	Dean Abramson	David Krauthemier
Notes As accurate as one can be at best this information should be used as *a guide* only. The manufacturer(s) should be directly contacted by the instrumentation specifying engineer in order to allow himself to best interpret the specifications on such a complex piece of equipment. Editor	(1A) 100 Hz with option 03; (1B) with option 02; (1C) Includes all options (Preselector-option 01)	(2A) Residual FM; (2B) At start of new sweep all drift eliminated; (2C) Frequency repeatability is assured by virtue of a synthesized l.o.	(3A) With 014 option; (3B) Using internal freq. adjust control accurate to ± 1 MHz; (3C) Using the internal calibrator.

Figure 1

fully programmable today via the IEEE-488 General Purpose Interface Bus (GPIB). This allows higher system test speeds and flexibility.

An example is complete two-way radio production testing where many parameters must be verified, requiring communications between spectrum analyzers, signal generators, frequency counters and multimeters.

Another example is RFI/EMI measurements which require a specific test procedure and various sensing devices

(Information supplied by "contributors")

Marconi TF2370 / TK2373	Polarad 640A-1
30 Hz to 1250 MHz	10 MHz to 40 GHz
20 Hz + 1% of span, width and ref.	± 5 MHz x n (5A)
(3 dB BW) 5 Hz to 50 kHz	(3 dB BW) 300 Hz to 1 MHz
NA	– 86 dBm including preselector
6 Hz P-P	150 Hz x n (5A)
100 dB	80 dB
100 Hz/10 minutes	± 5 kHz/10 minutes
75 dBc @ 30 kHz offset	>70 dBc @ 60 kHz offset
± 3 dB @ 1 GHz	± 1 dB to ± 5 dB (5B)
Data Out Only	IEEE-488 Compatibility For Digital Memory (5C)
Digital (4A)	Internal Calibrated Graticle
17" x 17-13/16" x 20-5/16"	8.75" x 16.75" x 17"
93 lbs.	55 lbs.
3.4	1.44
Not Specified	Not Specified
150 watts	180 watts
$39K (4C)	$15.25K (5D)
Keith Elkins	Ed Feldman

(4A) 130 x 100 mm television camera monitor tube-no storage CRT to burn out-no blooming
(4B) Has internal tracking generator which tracks the input tune frequency to within ± 2 Hz.
(4C) TF 2370: $19.5K
TK2373: $19.5K

(5A) n = mixing mode
(5B) Depending on control settings
(5C) Signal processing and signal storage functions
(5D) Includes option 1: internal digital memory and data interface.

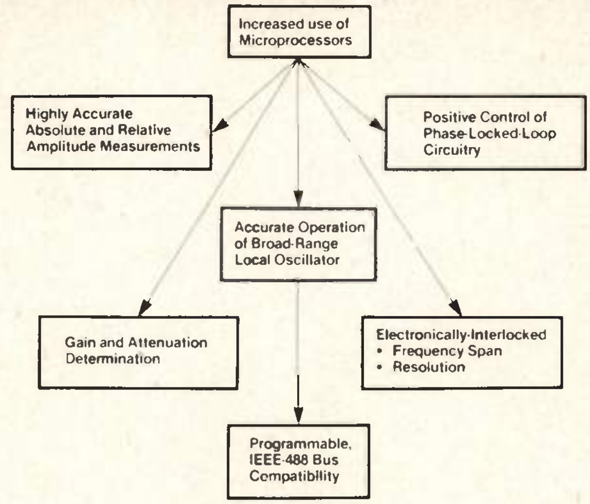


Figure 2

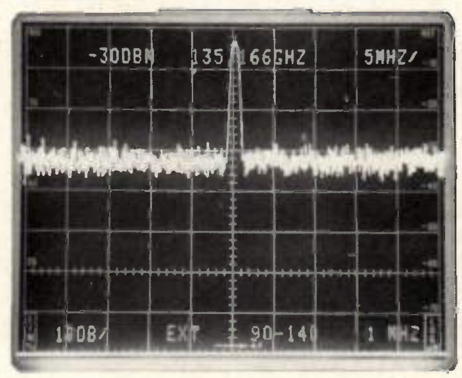


Figure 3

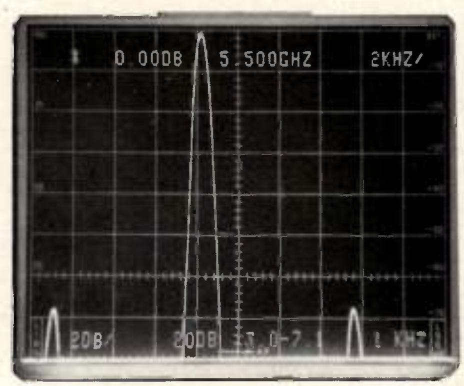


Figure 4 (a)

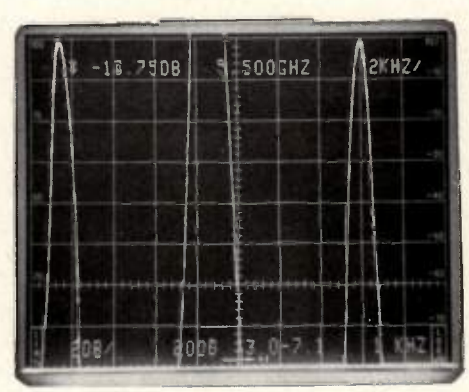


Figure 4 (b)

such as antennas and current probes. With microprocessor control over the bus, a step-by-step procedure implemented in software can issue commands and record data.

Rugged, Lightweight Packaging

Highly integrated electrical and mechanical design is exemplified by new millimeter wave mixer designs. A very small Schottky-barrier diode and probe across the high-field region of a ridged waveguide allows absolute amplitude measurements to 60 GHz.

Some spectrum analyzers are now available in all three basic physical configurations: portable, stackable bench-top, and rack-mount. Portable units[1] feature rugged construction for field use, are able to operate over a temperature range of −15 degrees C to +55 degrees C, and are designed to withstand moisture and a 12-inch drop (MIL T-28800B). Ruggedness results partially from a high shock-proof CRT, designed using computer simulation techniques.

Easier To Use

Modern spectrum analyzers may have as many as 30 different controls with the most frequently used being *center frequency, frequency span, resolution, gain, attenuation,* and *sweep time*. In many instances, center frequency is controlled by two or even three controls which are set for fast or slow tuning and for phaselocked or non-phaselocked operation, resulting in up to nine interacting controls. New instruments effectively reduce this number to as few as three.

By electronically interlocking frequency span, resolution, and sweep time, a microprocessor can determine appropriate settings with only one control knob being varied. Instead of the usual two or three controls, center frequency can be set by a single constant-tuning-rate (ctr) knob which controls frequency setting regardless of frequency span, while also maintaining the same signal movement on the screen whether at 500 MHz or 500 Hz per division. The disappearance of high-sensitivity controls that make the signal jump off the screen at the slightest touch is welcomed by users.

The newest innovation in control interlocking is the combination of attenuator and gain setting. Both of these traditional controls are replaced by two new functions: reference-level setting and minimum attenuation preset. The user simply presets the minimum attenuation that the measurement will tolerate as determined by the maximum power level, input VSWR needs, or other requirements. A single-knob reference level control is then used to set the level in either 10 dB or 1 dB steps; the microprocessor determining whether to insert gain or attenuation (See Figure 5).

The Future

Spectrum analyzers with GPIB compatibility will become more prominent in the years ahead to satisfy field needs of commercial, government and military users.

Technological advancements will include more higher performance microwave integrated circuits, more computing power per chip, and possibly flat-panel displays.

Higher accuracies are expected in areas such as absolute center frequency, where it can be expected that the spectrum analyzer will measure frequencies of displayed signals with counter accuracy. Amplitude measurement accuracy will likely increase by an order of magnitude from today's +1 dB to +0.1 dB, filling the need for a frequency selective power meter for pulsed and cw signals.

Footnote

1. Tektronix 492.

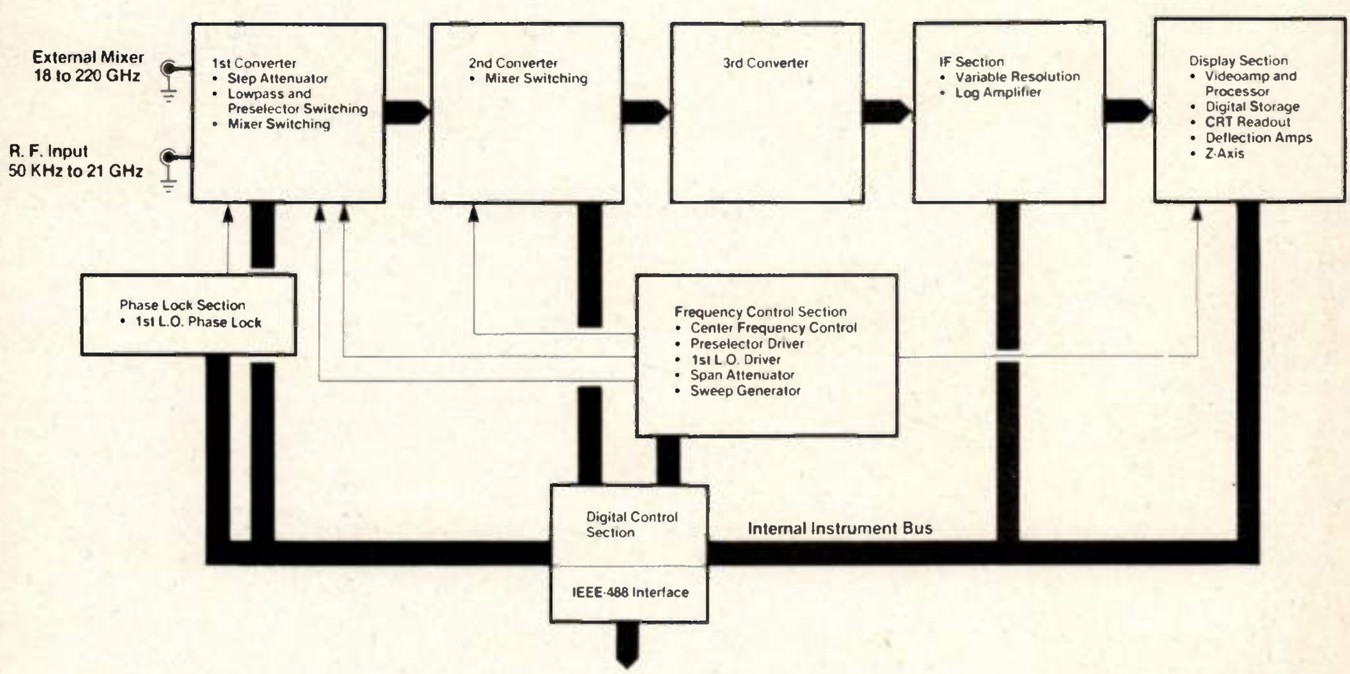
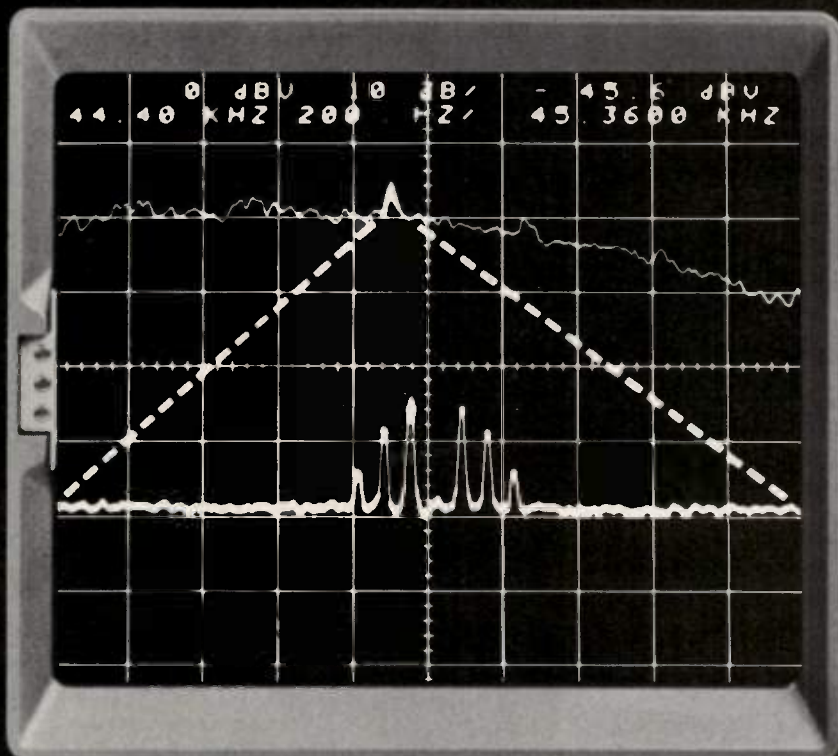


Figure 5

Increase the scope of your 'scope.



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INFO/CARD 5

S-Parameter Amplifier Design With Your HP-67

A two-card program that computes matching network component values from s-parameters.

Fred Hauer
EDMAC Associates Inc.
Rochester, N.Y.

Most VHF, UHF, and microwave transistors are characterized by scattering parameters. These are complex transmission and reflection coefficients measured at the input and output terminals of the device. Transmission coefficients may be either gain or attenuation, while reflection coefficients are directly related to VSWR and impedance. Being vector quantities, they contain magnitude as well as phase information and when used with a Smith Chart, can greatly simplify amplifier and matching network design. The four s-parameters are:

S_{11} The input reflection coefficient with the output port terminated in a matched load.
S_{22} The output reflection coefficient with the input port terminated in a matched load.
S_{21} The forward transmission gain with the output port terminated in a matched load.
S_{12} The reverse transmission gain with the input port terminated in a matched load.

An RF transistor amplifier can be designed quickly with the aid of the HP-67 calculator, using s-parameters. This article will not go into an elaborate discussion of s-parameters[1,2,3] but instead will describe a calculator program for amplifier design. This program lets the user design a transistor amplifier stage using s-parameters. The calculations include *Rollet's* stability factor, maximum gain, transducer input and output impedances and component values for input and output impedance matching to complex terminations. The matching networks are designed for a conjugate match to the input and output terminations. The choice of which matching network to use is based on practical considerations. Depending on the circuit application, these networks can be combined with biasing and filtering networks to reduce component count. If a particular type of network has to be used, and the component values are unrealistic, try matching to some intermediate image impedance. This is generally the geometric mean of the device impedance and the terminating impedance. Alternatively, the image can be selected to take advantage of the broadband matching afforded by 4:1 ratio toroidal transformers. Also, keep circuit Q low by selecting a network type where the first component *lowers* the SWR. This will result in the widest bandwidth and be less sensitive to parts value tolerances. In most cases you can go directly from calculator design to breadboard layout.

The routine begins with card #1 by computing Rollet's stability factor, K, which must be equal to or greater than 1 for an unconditionally stable amplifier. The closed form solution used in the remainder of the program is not valid if this factor is less than 1. The relationship used is:

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}|\,|S_{21}|}$$

where

$$\Delta = S_{11}S_{22} - S_{12}S_{21}$$

The maximum available power gain is then calculated:

$$A_{dB} = 10 \log \left|\frac{S_{21}}{S_{12}}\right| |K \pm \sqrt{K^2 - 1}|$$

The source and load reflection coefficients are found from:

$$\Gamma = C\left(\frac{B \pm \sqrt{B^2 - 4C^2}}{2|C|^2}\right)$$

where

$$C_s = S_{11} - \Delta S_{22} \qquad C_L = S_{22} - \Delta S_{11},$$

and

$$B_s = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2,\ B_L = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$$

These complex reflection coefficients are converted to the required series input and output impedances and are displayed in both polar and rectangular forms for convenience. It is not necessary to record them before loading card #2. However, it is important not to disturb the stack and to initialize immediately after loading card #2 to store data exactly as it is in the stack before doing anything else. Now input the frequency and the required terminating impedance. This impedance can obviously be the complex input or output impedance of another stage or it can be purely resistive — for a 50 ohm termination. Data entry format is always X followed by jY. Press E for input impedances and B for output impedances. Then select the impedance matching network that best fits your application.

Load Card #1

	Input	Key	Display		
	.32	Enter	320.0	– 03	
Store the S – parameters	111	CHS	– 111		
		A	320.0	– 03	
				– 03	
	.06	Enter	60.00	– 03	
	60	B	60.00	– 03	
	3.7	Enter	3.700	00	
	91	C	3.700	00	
	.71	Enter	710.0	– 03	
	23	CHS	– 23		
		D	710.0	– 03	
		E	1.055	00	= stability factor
Calc. gain		R/S	16.47	00	= gain (dB)
Display transducer input impedance			15.91	00	= Z in
			– 60.52	00	= Θ in
			7.829	00	= X in
			– 13.85	00	= jY in
Display transducer output impedance		R/S	189.1	00	= Z out
			– 76.88	00	= Θ out
			42.93	00	= X out
			– 184.2	00	= jY out

Load card #2

Input	Key	Display			
Initialize A		42.93	00		
frequency 400 x 10^6 fa		2.513	09	= 2πf	
Input impedance:					
X = 50	Enter	50.00	00		
jY = 0	E	50.00	00		
	C	0.000	00	no solution	(Type 1)
Calculate input network component values	f c	0.000	00	no solution	(Type 1)
	D	8.572	– 09	= E_1(Henrys)	(Type 2)
		– 92.07	– 12	= E_2(Farads)	(Type 2)
	f d	– 18.47	– 12	= E_1(Farads)	(Type 2)
		12.74	– 09	= E_2(Henrys)	(Type 2)
Output impedance:					
X = 50	Enter	50.00	00		
jY = 0	B	50.00	00		
Calculate output network component values	C	996.5	– 09	= E_1(Henrys)	(Type 1)
		78.73	– 09	= E_2(Henrys)	(Type 1)
	f c	40.19	– 09	= E_1(Henrys)	(Type 1)
		– 2.011	– 12	= E_2(Farads)	(Type 1)
	D	49.01	– 09	= E_1(Henrys)	(Type 2)
		66.36	– 09	= E_2(Henrys)	(Type 2)
	f d	– 3.230	– 12	= E_1(Farads)	(Type 2)
		80.22	– 09	= E_2(Henrys)	(Type 2)

There are 2 basic networks possible, each having 2 variations. The output is the shunt element E_1 followed by the series element E_2. Values are in Henrys and Farads with capacitors indicated by a minus sign. If 0.000 is displayed, that network is not valid — try another one. This part of the routine is based upon HP-67 program #EE1-02A with an error correcting routine and other changes.

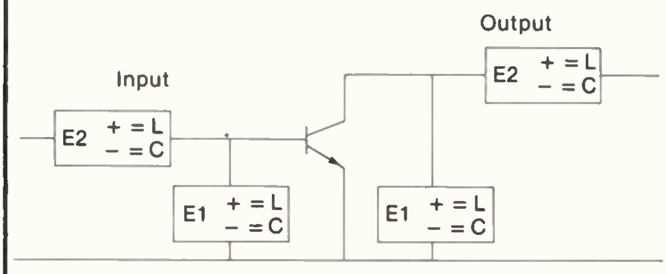
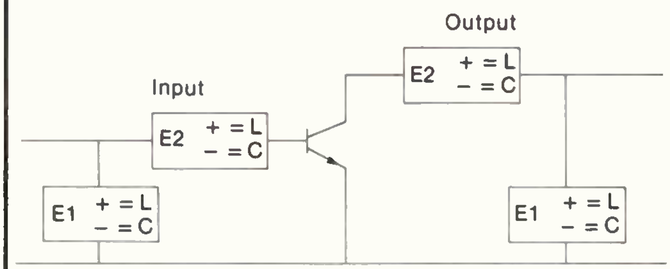


Figure 1A. Network C, c (Type 1).



A given design may have any type input network and any type output network. (Be sure to use the right layout for each.)

Figure 1B. Network D, d (Type 2).

Try the following example:

Given the following s-parameters for a 2N2857 biased at 5 ma. collector current at 400 MHz determine maximum power gain and L-C impedance matching to 50 ohms input and output.

S_{11} .32/– 111
S_{12} .06/60
S_{21} 3.7/91
S_{22} .71/– 23

Observe that there are only 2 possible solutions for the input network, while the output has four. Any of the output networks can be used with any of the input networks, so there is considerable design flexibility. The two types of networks are shown in Figure 1A and 1B. The shunt element, E_1, is nearest the transducer for Type 1 while the series element, E_2, is nearest the transducer for Type 2.

References

1. Perlow, S., "Transistor Amplifier Design Considerations," Microwave Journal, August 1977, pp. 22-30.
2. Martin, W.J., "HP-45 Calculator Speeds RF Amplifier Design," Electronics Magazine, March 20, 1975, pp. 134-135.
3. Hewlett-Packard Company, HP-67/HP-97 E.E. Pac. 1, Programs 02-01, 12-01, 16-01, 17-01, 18-01.

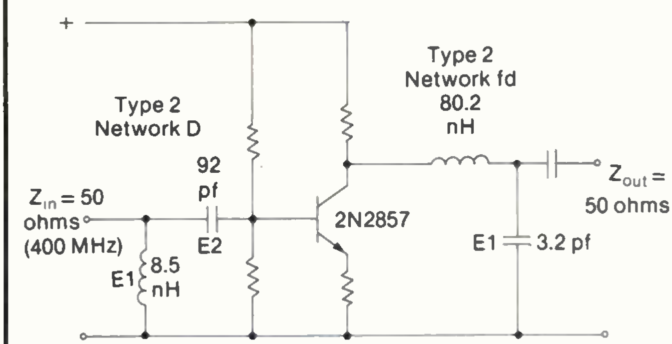


Figure 2. Amplifier Schematic Diagram.

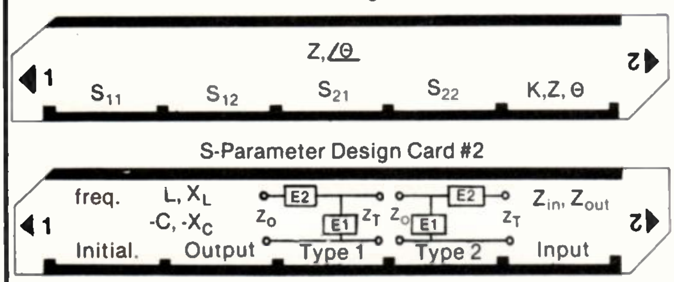


Figure 3. HP-67 Magnetic Card Labelling.

If, for some reason, you would like to recall the transducer input and output impedances calculated with card #1, key f e. The format here is:

X in =	7.829	00
jY in =	– 13.85	00
X out =	42.93	00
jY out =	– 184.2	00

The reactance rule included in the program is unique in that one key (f b) inputs inductance, capacitance, and reactance. The decision making process treats an input greater than 10^{-2} as a reactance. Anything less than 10^{-2} is capacitance in Farads or inductance in Henrys. A minus sign denotes capacitance and capacitive reactance. This covers the range of component values used in most RF designs. The frequency in Hz is entered using key f a. Load card #2.

Examples:

	Input	Key	Display		
Freq.	400×10^6	f a	2.513	09	$2\pi f$
L	10×10^{-9}	f b	25.13		Ohms X_L
X_L	50	f b	19.89	– 09	Henrys
X_C	– 50	f b	– 7.958	– 12	Farads
C	-15×10^{-12}	f b	– 26.53		Ohms X_C

A schematic diagram of the amplifier designed in the example is shown in Figure 2. Networks were selected to simplify DC biasing and to provide a low pass filter at the output.

Suggested labels for the two magnetic card HP-67 program are indicated in Figure 3.

The complete program listing for both cards are included in the Appendix.

CARD #1

Step	Key Entry	Key Code	Comments
001	f LBL A	31 25 11	
	STO 2	33 02	$= S_{11}\Theta$
	h x ≷y	35 52	
	STO 1	33 01	$= S_{11}Z$
	h RTN	35 22	
	f LBL B	31 25 12	
	STO 4	33 04	$= S_{12}\Theta$
	h x ≷y	35 52	
	STO 3	33 03	$= S_{12}Z$
010	h RTN	35 22	
	f LBL C	31 25 13	
	STO 6	33 06	$= S_{21}\Theta$
	h x ≷y	35 52	
	STO 5	33 05	$= S_{21}Z$
	h RTN	35 22	
	f LBL D	31 25 14	
	STO 8	33 08	$= S_{22}\Theta$
	h x ≷y	35 52	
	STO 7	33 07	$= S_{22}Z$
020	h RTN	35 22	
	f LBL E	31 25 15	
	h SF 1	35 51 01	
	RCL Σ+	34 21	
	h Σ−	35 21	
	RCL 2	34 02	
	RCL 8	34 08	
	+	61	
	RCL 1	34 01	
	RCL 7	34 07	
030	X	71	
	f →R	31 72	
	Σ+	21	
	RCL 4	34 04	
	RCL 6	34 06	
	+	61	
	RCL 3	34 03	
	RCL 5	34 05	
	X	71	
	f →R	31 72	
040	h Σ−	35 21	
	RCL Σ+	34 21	
	g →P	32 72	
	STO 9	33 09	$= \Delta_R$
	h x ≷y	35 52	
	STO 0	33 00	$= \Delta_\Theta$
	RCL 9	34 09	
	g X^2	32 54	
	1	01	
	+	61	
050	RCL 1	34 01	
	g X^2	32 54	
	−	51	
	RCL 7	34 07	
	g X^2	32 54	
	−	51	
	RCL 3	34 03	
	RCL 5	34 05	
	X	71	
	2	02	
060	X	71	
	÷	81	
	STO E	33 15	= K
	R/S	84	
	g X^2	32 54	
	1	01	
	−	51	
	f $\sqrt{X}$	31 54	
	RCL E	34 15	
	h ABS	35 64	
070	h x ≷y	35 52	
	−	51	
	h ABS	35 64	
	RCL 5	34 05	
	X	71	
	RCL 3	34 03	
	÷	81	
	f log	31 53	
	1	01	
	0	00	
080	X	71	
	f − X −	31 84	= Gain
	RCL Σ+	34 21	
	h Σ−	35 21	
	RCL 2	34 02	
	RCL 1	34 01	
	f →R	31 72	
	Σ+	21	
	RCL 0	34 00	
	RCL 8	34 08	
090	CHS	42	
	+	61	
	RCL 9	34 09	
	Enter	41	
	h R↓	35 53	
	RCL 7	34 07	
	X	71	
	f →R	31 72	
	h Σ−	35 21	
	RCL Σ+	34 21	
100	g →P	32 72	
	STO A	33 11	$= C_{SR}$
	h x ≷y	35 52	
	STO B	33 12	$= C_{S\Theta}$
	f GSB 4	31 22 04	
	GTO 5	22 05	
	f LBL 4	31 25 04	Calc B_s and B_L
	RCL 1	34 01	
	gX^2	32 54	
	RCL 7	34 07	
110	g X^2	32 54	
	h F ? 1	35 71 01	
	h X ≷y	35 52	
	h x ≷y	35 52	
	−	51	
	1	01	
	+	61	

Registers

0 $\Delta\Theta$	1 $S_{11}Z$	2 S_{11} Θ	3 $S_{12}Z$	4 $S_{12}\Theta$	5 $S_{21}Z$	6 $S_{21}\Theta$	7 $S_{22}Z$	8 $S_{22}\Theta$	9 Δ_R
S0	S1 X_s	S2 jYs	S3	S4	S5	S6	S7	S8	S9

A C_{SR}, C_{LR}	B $C_{S\Theta}, C_{L\Theta}$	C B_s, B_L	D Γ_s, Γ_L	E K	I

Step	Key Entry	Key Code	Comments
	RCL 9	34 09	
	g X^2	32 54	
	–	51	
120	STO C	33 13	$= B_s$ or B_L
	f x<0?	31 71	
	h SF2	35 51 02	
	h RTN	35 22	
	f LBL 5	31 25 05	
	f GSB 3	31 22 03	
	f GSB 0	31 22 00	
	R/S	84	
	GTO 2	22 02	
	f LBL 3	31 25 03	Calc $\lvert\Gamma_s\rvert$ or $\lvert\Gamma_L\rvert$
130	g X^2	32 54	
	4	04	
	RCL A	34 11	
	g X^2	32 54	
	X	71	
	–	51	
	f $\sqrt{X}$	31 54	
	RCL C	34 13	
	hX≷y	35 52	
	h F?2	35 71 02	
140	CHS	42	
	–	51	
	RCL A	34 11	
	g X^2	32 54	
	2	02	
	X	71	
	÷	81	
	RCL A	34 11	
	X	71	
	STO D	33 14	$= \lvert\Gamma_s\rvert$ or $\lvert\Gamma_L\rvert$
150	h RTN	35 22	
	f LBL 0	31 25 00	Calc Z/Θ
	RCL B	34 12	
	Enter	41	
	RCL D	34 14	
	1	01	
	Enter	41	
	h R↓	35 53	
	h R↓	35 53	
	f →R	31 72	
160	h R↑	35 54	
	–	51	
	h STI	35 33	
	h R↓	35 53	
	h R↓	35 53	
	+	61	
	+	61	
	g →P	32 72	
	h R↑	35 54	
	h RCI	35 34	
170	g →P	32 72	

Step	Key Entry	Key Code	Comments
	h R↓	35 53	
	hx≷y	35 52	
	h R↑	35 54	
	÷	81	
	h R↓	35 53	
	–	51	
	hR↑	35 54	
	5	05	
	0	00	
180	CHS	42	
	X	71	
	f →R	31 72	
	g →P	32 72	
	fP≷S	31 42	
	f –X–	31 84	$= \lvert Z\rvert$
	hx≷ y	35 52	
	f –X–	31 84	= /Θ
	hx≷y	35 52	
	f →R	31 72	
190	h F?1	35 71 01	Source or load?
	STO 1	33 01	$= X_s$ (Store Source)
	f –X–	31 84	
	hx≷y	35 52	
	h F?1	35 71 01	Source or Load?
	STO2	33 02	= jYs (Store source)
	fP≷S	31 42	
	h RTN	35 22	
	f LBL 2	31 25 02	Repeat for Load
	h CF 1	35 61 01	
200	RCLΣ+	34 21	
	h Σ–	35 21	
	RCL 8	34 08	
	RCL 7	34 07	
	f →R	31 72	
	Σ+	21	
	RCL 0	34 00	
	RCL 2	34 02	
	CHS	42	
	+	61	
210	RCL 9	34 09	
	Enter	41	
	h R↓	35 53	
	RCL 1	34 01	
	X	71	
	f →R	31 72	
	h Σ–	35 21	
	RCL Σ+	34 21	
	g →P	32 72	
	STO A	33 11	$= C_L R$
220	hx≷y	35 52	
	STO B	33 12	$= C_L \Theta$
	f GSB 4	31 22 04	
	f GSB3	31 22 03	
	GTO 0	22 00	

Labels

A	B	C	D	E
A Store S_{11}	B Store S_{12}	C Store S_{21}	D Store S_{22}	E Calc. K,A,Z/Θ
a	b	c	d	e used
0 Calc Z/K Θ	1	2 Repeat	3 Calc r	4 Calc B
5 used	6	7	8	9

Flags

Flags
0
1 used
2 used
3

Set Status

Flags	ON	OFF	Trig		Disp	
0	☐	X	Deg	X	Fix	☐
1	☐	X	Grad	☐	Sci	☐
2	☐	X	Rad	☐	Eng	X
3	☐	X			n 3	

CARD #2

Step	Key Entry	Key Code	Comments
001	f LBL A	31 25 11	Initiate
	STO 2	33 02	= jYL
	h x ≷y	35 52	
	STO 1	33 01	= X_L
	h RTN	35 22	
	f LBL B	31 25 12	Output Network
	h CF 0	35 61 00	
	h CF 1	35 61 01	
	h CF 2	35 61 02	
010	STO 4	33 04	Store jY out
	h R↓	35 53	
	STO 3	33 03	Store X out
	h RTN	35 22	
	f LBL E	31 25 15	Input Network
	h CF 0	35 61 00	
	h SF 1	35 51 01	
	h CF 2	35 61 02	
	f p ≷s	31 42	Secondaries
	STO 4	33 04	Store jY in
020	h R↓	35 53	
	STO 3	33 03	Store X in
	f p ≷s	31 42	
	h RTN	35 22	
	f LBL 1	31 25 01	Calc E_1
	RCL 2	34 02	
	g X^2	32 54	
	RCL 1	34 01	
	g X^2	32 54	
	+	61	
030	RCL 3	34 03	
	X	71	
	RCL 1	34 01	
	RCL 3	34 03	
	−	51	
	÷	81	
	h LST x	35 82	
	h 1/x	35 62	
	RCL 3	34 03	
	X	71	
040	RCL 2	34 02	
	X	71	
	g X^2	32 54	
	h LST x	35 82	
	h R↓	35 53	
	+	61	
	f x<0?	31 71	Test
	GTO 7	22 07	= Not valid
	f $\sqrt{X}$	31 54	
	h R↑	35 54	
050	h x ≷y	35 52	
	−	51	
	STO 5	33 05	
	h LST X	35 82	
	Enter	41	
	+	61	

Step	Key Entry	Key Code	Comments
	+	61	
	STO 6	33 06	
	h RTN	35 22	
	f LBL C	31 25 13	Type 1 Network
060	h F?1	35 71 01	
	f p ≷s	31 42	
	f GSB 1	31 22 01	
	STO 9	33 09	
	f GSB 5	31 22 05	
	RCL 9	34 09	
	f GSB 3	31 22 03	
	f GSB 5	31 22 05	
	h F?1	35 71 01	
	f p≷s	31 42	
070	h RTN	35 22	
	g LBL f c	32 25 13	Type 1' Network
	h F?1	35 71 01	
	f p≷s	31 42	
	f GSB 1	31 22 01	
	RCL 5	34 05	
	STO 9	33 09	
	f GSB 5	31 22 05	
	RCL 9	34 09	
	f GSB 3	31 22 03	
080	f GSB 5	31 22 05	
	h F?1	35 71 01	
	f p≷s	31 42	
	h RTN	35 22	
	f LBL D	31 25 14	Type 2 Network
	h SF 2	35 51 02	
	h F?1	35 71 01	
	f p≷s	31 42	Exchange if Input
	f GSB 2	31 22 02	
	f GSB 1	31 22 01	
090	STO 9	33 09	
	f GSB 5	31 22 05	
	RCL 9	34 09	
	f GSB 3	31 22 03	
	f GSB 5	31 22 05	
	+	61	
	f GSB 2	31 22 02	
	h LST x	35 82	
	h F?1	35 71 01	
	f p≷s	31 42	
100	h CF2	35 61 02	
	h RTN	35 22	
	g LBLfd	32 25 14	Type 2' Network
	h SF 2	35 51 02	
	h F?1	35 71 01	
	f p≷s	31 42	
	f GSB 2	31 22 02	
	f GSB 1	31 22 01	
	RCL 5	34 05	
110	f GSB 05	31 22 05	
	RCL 9	34 09	

Registers

0 $2\pi f$	1 X_L	2 jY_L	3 X_{out}	4 jY_{out}	5 used	6 used	7 used	8	9 used
S0 $2\pi f$	S1 X_s	S2 jYs	S3 X_{in}	S4 jY_{in}	S5 used	S6 used	S7 used	S8	S9 used
A	B	C	D	E	I				

Step	Key Entry	Key Code	Comments
	f GSB 5	31 22 05	
	+	61	
	f GSB 2	31 22 02	
	h LST x	35 82	
	h F? 1	35 71 01	
	f p≷s	31 42	
	h CF 2	35 61 02	
120	h RTN	35 22	
	f LBL 3	31 25 03	Calc E_2
	STO 7	33 07	
	RCL 2	34 02	
	+	61	
	RCL 3	34 03	
	X	71	
	RCL 7	34 07	
	RCL 4	34 04	
	+	61	
130	RCL 1	34 01	
	X	71	
	−	51	
	RCL 1	34 01	
	÷	81	
	h RTN	35 22	
	f LBL 2	31 25 02	Exchange Registers
	RCL 1	34 01	
	RCL 3	34 03	
	STO 1	33 01	
140	h x≷y	35 52	
	STO 3	33 03	
	RCL 2	34 02	
	RCL 4	34 04	
	STO 2	33 02	
	h x≷y	35 52	
	STO 4	33 04	
	h RTN	35 22	
	f LBL 5	31 25 05	
	h F? 0	35 71 00	
150	GTO 6	22 06	
	h SF 0	35 51 00	
	f GSB 9	31 22 09	
	h RTN	35 22	
	f LBL 6	31 25 06	
	h CF 0	35 61 00	
	f LBL 9	31 25 09	Calculate, L,C
	f x<0?	31 71	X_L or X_c?
	GTO 4	22 04	if X_c
	RCL 0	34 00	Calculate L
160	÷	81	
	f −X−	31 84	Display
	h RTN	35 22	
	f LBL 4	31 25 04	Calculate C
	RCL 0	34 00	
	X	71	
	h 1/X	35 62	
	f −X−	31 84	Display
	h RTN	35 22	
	R/S	84	
170	f LBL 7	31 25 07	Error Correcting Routine
	h F?2	35 71 02	
	f GSB 2	31 22 02	
	h F? 1	35 71 01	
	f p≷s	31 42	
	0	00	
	R/S	84	
	g LBL fa	32 25 11	Store frequency and calculate ω
	hπ	35 73	
	X	71	
180	2	02	
	X	71	
	STO 0	33 00	
	f p≷s	31 42	
	STO 0	33 00	
	f p≷s	31 42	
	h RTN	35 22	
	R/S	84	
	g LBLF e	32 25 15	Recall Z_{in}, Z_{out}
	f p≷s	31 42	
190	RCL 1	34 01	X_s
	RCL 2	34 02	jYs
	f p≷s	31 42	
	RCL 1	34 01	X_L
	RCL 2	34 02	jY_L
	g STK	32 84	
	h RTN	35 22	
	g LBL f b	32 25 12	Reactance Rule Routine
	h CF2	35 61 02	
	f x<0?	31 71	
200	h SF2	35 51 02	
	h ABS	35 64	
	EEX	43	
	2	02	
	CHS	42	
	g x>y?	32 81	
	GTO 8	22 08	
	h x≷y	35 52	
	h F?2	35 71 02	
	CHS	42	
210	GTO 9	22 09	
	f LBL 8	31 25 08	
	h x≷y	35 52	
	h F?2	35 71 02	
	GTO 8	22 08	
	RCL 0	34 00	
	X	71	
	h RTN	35 22	
	f LBL 8	31 25 08	
	RCL 0	34 00	
220	X	71	
	h 1/x	35 62	
	CHS	42	
	h RTN	35 22	
	R/S	84	

Labels

A Initiate	B Output	C Type 1	D Type 2	E Input
a Freq	b L,C X_L, X_c	c Type 1'	d Type 2'	e Recall X + Jy
0	1 used	2 used	3 used	4 used
5 used	6 used	7 used	8 used	9 used

Flags

0 used
1 used
2 used
3

Set Status

Flags	ON	OFF	Trig		Disp	
0	☐	☒	Deg	☒	Fix	☐
1	☐	☒	Grad	☐	Sci	☐
2	☐	☒	Rad	☐	Eng	☒
3	☐	☒			n 3	

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EMI and RFI — Some Causes and Cures

A broad sweep review of the causes and techniques for reducing RFI/EMI.

Howard Walker
Pace, Inc.
Silver Springs, MD

EMI Defined

Technically, EMI is, "any Electrical or Electromagnetic phenomenon — Man Made or natural — causing an undesirable response, performance degradation, or complete malfunction of operational electronic equipment. EMI varies, in time and degree, from a nuisance to complete destruction of mission performance."[1]

The sources of EMI are seemingly unlimited. In fact, any electromechanical, electric or electronic device is a possible source of EMI. Even mother nature is sometimes the culprit via atmospheric and cosmic disturbances.

3 Major Causes of EMI

- **Natural Sources** — Generally due to lightning.
- **Galactic** — Originates outside the earth's atmosphere. Caused by solar flares of our sun or other stars millions of miles away. This noise is usually found between 18 and 500 MHz.
- **Man-Made** — Due mostly to motors, power lines and transformers, neon signs, vehicle ignition, and medical and industrial equipment. This type mainly occurs below 20 MHz.

Man-made EMI sources include equipment whose function is to intentionally generate or radiate electromagnetic signals, and equipment that unintentionally generates electromagnetic energy. Unfortunately, even when electromagnetic energy is intentionally generated and presumably restricted to the fundamental or intended frequency range, it sometimes generates harmonic frequencies that cause interference problems with other equipment.

EMI has been a factor since the first electromagnetic wave was propagated intentionally. Also known as Radio Noise, Electrical Noise, and Radio Frequency Interference (RFI), it covers the spectrum from DC to light frequencies.

Until recently, users and manufacturers of electrical and electronic equipment were able to ignore the EMI generated by their equipment. Now, however, more stringent regulations make manufacturers and users alike responsible for interference caused by their equipment.

More stringent regulations have been necessitated, not only by the abundance of equipment crowding the spectrum, but also by the nature of the modern developments. The increased use of semiconductors and integrated circuits, as well as the increased use of plastics in cabinets and housings; instead of metal, have caused a problem with EMI leakage into, as well as, out of equipment.

Plastics have many advantages, but like all insulators, they are poor EMI shields; without special conductivity treatment. Although susceptibility to EMI may be reduced by filtering, this must be augmented with good design of housing and cabinets. The housing should act as a two-way shield, either keeping EMI out; or in the case of an EMI generating device; keep it in.

The ultimate goal of the various EMI specifications is to ensure Electromagnetic Compatibility (EMC) — that equipment and systems will fucntion as designed without degradation of malfunction in their intended operational electromagnetic environment nor adversely affect the operation of any other equipment or system.

The FCC has the responsibility of controlling the generation of any electrical interference that hampers communications.

The FCC is only one of the many government agencies involved in regulating policy and controlling EMI. The other primary agencies are:

- Office of Telecommunications Policy (OTP)
- Office of Telecommunication
- Department of Commerce
- Department of Defense (DOD) and it's Tri-Service Executive Agencies
- National Aeronautics and Space Administration (NASA)
- Department of Transportation (DOT) including it's Federal Aviation Agency (FAA)
- Bureau of Radiological Health (BRH)
- The National Security Agency (NSA)

In addition, there are many others such as the National Bureau of Standards and the Environmental Sciences and Service Administration, which play an important role in EMI, but are not responsible for issuing EMI specifications, rules, regulations and policy.

Eliminating EMI

Equipment redesign is not cost effective and, often times, not EMI effective. Prevention seems to be the most practical procedure, both in terms of economics and results. Grounding, shielding, balanced lines, twisted pairs and filtering methods are used to prevent or contain EMI.

- **Grounding**

A proper ground, for interference reduction, should be a zero impedance point which serves as a single reference point. Since no two points are at the same electrical potential, multiple ground connections introduce a "ground loop", which conducts interference to other parts of the system.

Grounding should not be considered a cure-all, however. In some cases it may be necessary to be used *with* shielding. But, *grounding is not shielding*. The housing sections may be grounded, but still leak (radiation).

- **Shielding**

Electric (E) and Magnetic (H) fields travel through free space at 3×10^8 meters/second. When these waves strike a conductor, some energy is reflected and the incident wave also sets up eddy currents in the conductor, which attenuate the signal. Thus, a shield reduces the effect of the electromagnetic wave by reflection and energy absorption.

For electrically thick material:

$$R = 50 + 10 \log_{10}(PBf)^{-1} \qquad (1)$$

$$A = 1.7t\,(f/PB)^{.5} \qquad (2)$$

Where R = Reflected Wave in dB
A = Absorbed Wave in dB
PB = in ohms/cm
f = Frequency in MHz
t = Thickness in cm

A is dependent on frequency, thickness, conductivity and permeability of the shield material.

A major problem with shielding is that any discontinuity in the shield can leak EMI.

Shielding in plastic cabinets can be achieved using:

Conductive Foil — Generally applied inside the cabinet. It usually requires a coating to prevent corrosion. Effectiveness of 55-100 dB for 5 mil thickness, is achieveable.

Conductive Coatings — Also applied inside cabinets. Coatings include graphite, silver, nickel and zinc. The first three are applied with an organic solvent carrier. Arc and flame spraying techniques are used to apply zinc by utilizing an electric current or gas flame to melt zinc or a zinc alloy wire. Droplets of the metal are then blown onto the housing suface through an air jet.

Conductive Plastics — Thermoset and Thermoplastic resins are made conductive with the addition of metal flakes or fibers, carbon powder or graphic fibers.

- **Shielding for RFI**

Shielding is perhaps the most important method of preventing or curing RFI. Maximum effects of a particular type shield are realized with no breaks or points of entry (in the shielding).

Ventilation Holes — Honeycomb type vent holes are effective in that the small tubing of the honeycomb act as a waveguide below cutoff.

Coaxial Cable — Proper termination is a must. The braid should be soldered so that the inner conductor is surrounded by the braid at the termination point.

Improperly designed openings in keyboards or ports, and faulty welding can act as waveguide antennas. Above 100 MHz, leakage is serious, as the required slot size becomes smaller. The slot's length, not the width or area, determine it's shielding effectiveness. For slot lengths smaller than the interference wavelength an enclosure will shield. However, as the frequency increased, shielding effectiveness reaches a minimum at the slot resonant frequency.

Panels must be in full contact with the body of the housing. A conductive gasket (i.e. braid or graphite impregnated rubber) must be used. In the case of two or more major housing sections, self-tapping screws are wise to use to attain maximum conductivity.

- **Coaxial Cable**

Provides an effective means of shielding an interconnecting line against interference and reducing radiation from the line.

- **Balanced Lines**

This method to minimize EMI susceptibility utilizes a *balun* transformer at the source and load. Primary and secondary windings are shielded from each other and the transformers are usually grounded. Balanced lines minimize induced voltages and currents caused by interfering fields and interference caused by a difference in ground potential.

- **Twisted Pairs**

Twisting the conductors along the length of the cable provides signal cancellation of EMI, but, the length of twist is difficult to control. This method is only effective when the EMI frequency wavelength is shorter than the length of the twist. This method is not recommended for use in circuits with multiple grounds.

- **Filters**

EMI can enter equipment through power and signal lines, or through the enclosure via radiation. The previously mentioned techniques are employed to minimize radiated interference, but they are ineffective against conducted interference. EMI filters on power or signal lines are then called for.

Power line filters, which are usually *Low Pass*, are designed to pass frequencies up to the low kHz range.

Filters used on signal and power lines are classified by their type of frequency discrimination.

- **Low Pass** — Pass frequencies from DC to a specified cutoff frequency and attenuate all signals above by an increasing amount.
- **High Pass** — Pass signals above a specified cutoff frequency and attenuate all signals below by an increasing amount.
- **Band Pass** — Pass all signals in a specified band and attenuate all signals above or below cutoff.
- **Band Reject** — Pass all frequencies except those in a specified band.

Chebyshev & Elliptic Function Filters for RFI

The Chebyshev low-pass, a common type of filter useful for this application consists of alternating series L and shunt C components. The attenuation[5] of the filter is specified by

$$L\,(dB) = 10 \log_{10} \frac{1}{1 + \omega\,\varepsilon^2\, T_n^2(\omega)} \qquad (3)$$

where

$\omega = 2\pi f$ (*f is the frequency in Hertz*)

ε^2 = *ripple factor (as a numeric)*

$T_n(\omega)$ = *Chebyshev polynomial of degree n* and where

$$\varepsilon^2 = 10^{\frac{ripple\,(dB)}{10}} - 1$$

n = total number of L(s) and C(s)

$$T_n(\omega) = \begin{cases} \cos(n \arccos \omega) & 0 \leq \omega \leq 1 \\ \cosh(n \operatorname{arccosh} \omega) & \omega > 1 \end{cases}$$

A more complex and more (characteristic) controllable filter, similar in construction to the Chebyshev, is the Elliptic Function filter. The shunt C(s), of the Chebyshev, are replaced by series L(s) and C(s). This has the advantage of providing a sharper roll-off with increasing frequency (past the 3 dB or half-power point).

Filter selection involves several factors, the first of which are:

Frequency of Operation

The filter selected must have within it's range the frequencies to be filtered.

Working Voltage

The voltage level of the application must be within the working voltage specification. Most *commercial* filters are rated in the 125-200 VDC range and a few in the 500-600 VDC range.

Insertion Loss

To accurately determine the amount of insertion loss, impedances within the circuit must be known. Since this is not normally available, a filter is chosen with as high an insertion loss/volume ratio as possible.

Current Rating

The DC current load rating should be 1.5 to 3 times the expected application.

RF Current

The type of filter construction limits the amount of RF power it can dissipate. Some filters incorporate capacitors that will not withstand much RF current, while others convert the RF current to heat. RF current values in most applications are less than .25 amps.

Insulation Resistance

This is a function of the dielectric material used in the filter. This parameter determines the leakage current.

Dielectric Voltage Test

This is a short term test (5-60 seconds) that stresses the filter's dielectric materials 2-5 times the working voltage level.

D.C. Resistance

This is a function of the length and diameter of the wire used. Basically, it is the resistance of the wire, and it becomes important only in very high current applications where voltage drop is a concern.

Operating Temperature Range

Industry standards are normally −55°C to +125°C for military and −25°C to +85°C for commercial type applications.

Storage Temperature

Storage temperature is sometimes overlooked. Users often store and transport equipment in environments more severe than the equipment encounters in actual operation.

Noise Generation

Theremal noise is caused by the agitation of electrons in resistance. The mean square value of thermal noise (E^2) is:

$$E^2 = 4\,RKT \cdot \Delta f$$

Where

K = Boltzmann's Constant (1.38×10^{-23} joules/°K)
T = Absolute Temperature (in degrees Kelvin)
Δf = Bandwidth (Hz.)
R = Resistance (in ohms) or real component of Z

Noise Measurement

RFI and EMI are measured in the same manner as radio wave field strength. However, peak rather than average values of noise are generally of interest. The overall bandpass of the measuring device must be accurately known.

Calculation of Noise Values

Estimation of noise levels in a receiver, due to external noise, requires taking into account two factors:

- Orientation of the receiving antenna.
- Gain of receiving antenna.

In general, noise power is proportional to bandwidth. Thus, where

F_a = Effective Antenna Noise Factor

Then

$$F_a = P_n/KT_oB = T_a/T_o$$

Where P_n = Noise power from equivalent lossless antenna (Watts)
K = Boltzmann's Constant
T_o = Reference Temperature (290° Kelvin)
B = Effective Receiver Noise Bandwidth (Hz)
T_a = Effective antenna temperature with external noise (Kelvin)

References

1. Reference Data for Radio Engineers, Radio Noise and Interference, Atmospheric Noise, Howard W. Sams and Co.
2. CCIR Report 322 10th Plenary Assembly, Geneva; 1963.
3. "EMI Filters", Hopkins Engineering Co. San Fernando, Calif.
4. "AMP Quiet Line Filter Handbook", AMP Incorporated Harrisburg, PA.
5. The Radio Amateur's Handbook. American Radio Relay League. Newington, CT 1979.

Miniature Rotary Attenuator DC-1 GHz

Texscan Corporation announces the addition of its MA-211 attenuator. The new product offers 3 watt average power handling, DC-1 GHz frequency coverage and minimal package density. It provides 0-60 dB in 10 dB step attenuation. Weighing only 6 1/2 oz. and packaged in a case diameter measuring 1.0 inches, the MA-211 occupies minimum behind-the-panel space and is constructed to offer superior RFI performance.

Specifications include attenuation, 0 to 60 dB in 10 dB steps; impedance, 50 ohms; frequency range, DC to 1 GHz; accuracy, ±1 percent or 0.3 dB at 30 MHz, ±2 percent or 0.3 dB at 500 MHz, ±3 percent at 1000 MHz; inserstion loss, 0.2 dB; maximum VSWR, 1.2:1 at 500 MHz, 1.3:1 at 1 GHz; average power handling capability, 3 W; peak power handling capability, 100W, 3μ Sec. Pulse; operating temperature range, 0° to +55°C; temperature stability, 1 x 10^{-4}; rotation angle between steps, 30°; connectors, SMA; weight, 6 1/2 oz. (185g.).

Contact: Texscan Corporation, 2446 North Shadeland Avenue, Indianapolis, Indiana 46219. (317) 357-8781. Circle INFO/CARD #126.

400 MHz Equipment

C-COR's 400 MHz equipment will include trunk amplifiers with an operational spacing of 22 dB and distribution amplifiers similar in configuration to C-COR's D-500 series with a bandpass of 54-400 MHz.

This equipment will have two-way capability and the reverse system will have a bandpass of 5-30 MHz. All main line passives will have a frequency response from 5 to 400 MHz. Shipment of the equipment will depend upon the availability of 400 MHz hybrids from the hybrid manufacturers. The performance of the equipment will be a function of those hybrids.

Contact C-COR Electronics, Inc., (814) 238-2461. INFO/CARD #125.

Front-End Crystal Filters

IXF front-end crystal filters for VHF high-band and low-band operation offer protection against overloading and intermodulation. Two-pole and four-pole models allow you to choose the selectivity you need. Data sheets available on request.

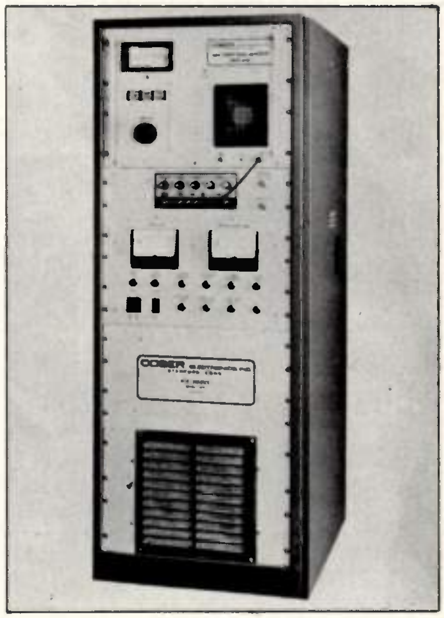
Contact Piezo Technology Inc., P.O. Box 7859, Orlando, Fla. 32854. (305) 298-2000. INFO/CARD #118.

High Power Microwave Signal Generator

Cober Electronics announces a new line of pulsed high power signal generators for the microwave industry. Designated the series 1691, these units satisfy the need for an economical, precisely controllable, laboratory microwave transmitter. The system provides peak powers continuously adjustable from 200 watts to 1.8 KW over the frequency range of 975 MHz to 10,475 MHz using 8 interchangeable heads. A system may be ordered with a full complement of RF heads or individually for the particular frequency range of interest at the time. Additional heads can be ordered as frequency requirements change. Pulse width is continuously variable from 1 to 10 microseconds at duty cycles of up to 10 percent. Other pulse width ranges are available as options. Complete metering and monitoring are included as well as protection for the signal generator, load, and personnel.

The series 1691 is used for the high power testing of microwave components such as filters, RF assemblies, circulators and attenuators, for RFI and EMC measurements, and for plotting antenna patterns. The system is also suitable for use as a microwave source for studying the effects or potential hazards of high power microwave radiation. Other applications include use for the evaluation of radar systems and as an ECM simulator. The model 1691 is 21"W x 27"D x 60"H and weighs approximately 600 lbs. Delivery is 120-150 days.

Contact Coberg Electronics, Inc., 7 Gleason Ave., Stamford, Conn. 06902. INFO/CARD #123.



Broadband Linear Amplifier

ENI, Inc. announces the model 3200L RF broadband, solid state power amplifer. With a linear power level exceeding 175 watts over a frequency of 250 kHz to 150 MHz, the model 3200L is a compact and efficient power amplifier. This linear Class A unit has a high gain of 55 dB making it compatible with any commercial synthesizer, signal or sweep generator. The wide frequency and linear phase response of this instrument enables it to amplify AM, FM, SSB, TV or pulse signals with minimum distortion. Unconditional stability and instantaneous failsafe provisions in the unit provide absolute protection from damage due to transient and overload conditions. It will operate continuously into any load impedance from an open to a short circuit without causing damage or oscillation.

For applications where slightly higher distortion levels can be tolerated the model 3200L will provide more than 250 watts of CW and pulse power from 2 to 80 MHz. This unit is complete with integral power supply and cooling system for operation from 115/230 VAC. The wide range of applications for the model 3200L include linear accelerator drive, RFI/EMI susceptibility testing, on-the-air transmission, plasma generation, power meter calibration, NMR/ENDOR spectroscopy and general laboratory instrumentation.

Contact Electronic Navigation Industries, Inc., 3000 Winton Rd. South Rochester, N.Y. 14623. (716) 473-6900. INFO/CARD #111.

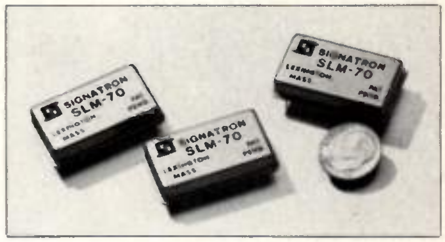
Linear RF Multiplier

A true linear RF multiplier for weighting RF signals in adaptive systems is being introduced by Signatron, Inc. of Lexington, Massachusetts.

The SIGNATRON model SLM-70 is a true linear RF multiplier that differs from double balanced mixers by providing an output signal level linearly proportional to the signal levels at both inputs. RF input range is 40 to 200 MHz; baseband input range, DC to 230 kHz; voltage driven baseband input, ±10 VDC; linearity, ±4 percent; dynamic range, 40 dB; and 3rd order intercept, +30 dBm.

Hermetically sealed in a 1.37"L x 0.80"W x 0.29"H dual in-line package, the 0.5 oz. SIGNATRON model SLM-70 operates over a 0 to 50°C temperature range or −25° to +75° with slightly degraded specifications. Power requirement is +15 VDC. Typical applications include complex signal weighting, phase shifter circuits, modulation, amplitude modulation, and voltage controlled linear attenuators.

Contact Signatron, Inc., 12 Hartwell, Avenue, Lexington, Mass. 02173. (617) 861-1500. INFO/CARD #120.

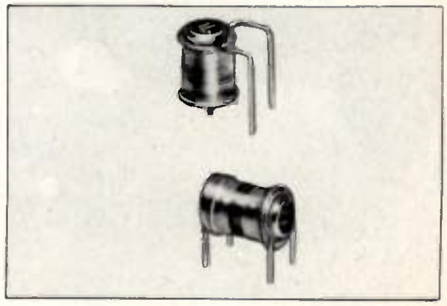


Precision Trimmer Capacitors

The SF/SP line of precision trimmer capacitors has been expanded by Voltronics to include many more tuning ranges formerly covered only by the larger TF/TP styles. All of these sealed glass dielectric capacitors use Voltronics non-rotating piston design. Temperature coefficients are ±50 or ±150 PPM/°C. Seal withstands 40 PSI. Typically, a 1 to 20 pF SP20, a horizontally mounted, .440" long P.C. style, is 40 percent shorter and has 25 percent more capacitance than rotating piston MIL-C-14409D types.

New avionic, communication and military devices such as receivers using synthesizer techniques need only one critical precision trimmer capacitor where old designs had from 5 to 15. The SF/SP line designed for these applications is not only small and completely sealed, but also has the highest resolution of any trimmer — 72 turns per inch. Its tuning is linear without reversals because of the non-rotating piston action. The line has been approved in many new OEM designs using these types of circuits. It is now available in standard maximum capacitance ratings of 4.5, 5.5, 8.5, 10, 11, 12, 14, 16, 17, 20, 22, 25, 28, 30, 36 and 40 pF in both horizontal and vertical P.C. mounting.

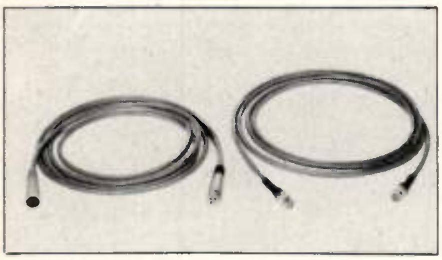
Contact Voltronics, West Street, East Hanover, N.J. 07936. (201) 887-1517. INFO/CARD #106.



Armored Cable Assemblies

Trompeter Electronics has developed new armored coax, audio and electro optic cable assemblies that combines the excellence of our Wrench-Crimp BNC connector with the physical protection and supplemental R-F and hum shielding of a flexible stainless steel sheath. Intended for abusive environments such as television electronic news gathering and military field use, these rodent-proof cables withstand pull loads well in excess of 150 pounds and are unaffected by crush loads associated with heavy automotive and industrial traffic. Coax assemblies can be fabricated from either 50 ohm (RG122) or 75 ohm (Belden 8218) miniature cable; audio assemblies from Trompeter Twinax miniature cable; Electro Optic assemblies from a variety of fiber cable from 300μm to 215μm. Custom cable assemblies for various uses are available on special order. These new cable assemblies are available in standard lengths of 10 to 50 feet in 10 foot increments, or custom lengths to your special order. A free catalog listing the many patching systems and coax connectors is available.

Contact Trompeter Electronics, Inc., 8936 Comanche Avenue, Chatsworth, Calif. 91311. (213) 882-1020. Circle INFO/CARD #119.



A Versatile DC To 500 MHz Amplifier

The CLC100 amplifier combines wide bandwidth, very flat passband response, low distortion, and excellent pulse response into one unit. A 500 MHz-3 dB bandwidth, as well as passband flatness of ±.15 dB max from DC to 300 MHz are obtained while peaking is held to +.2 dB max. The deviation of the phase response from linear is 5° max to 300 MHz. Total harmonic distortion is less than .3 percent from DC to 200 MHz, at 0 dBm, while the two-tone third-order intercept is +27 dBm at 200 MHz. The rise time is 750 ps max and settling time to 2 percent is 10 ns max. Nominal gain is 19.6 dB while the −1 dB power output compression level is +12 dBm. Input and output impedances are 50 ohms. Supply current is 45 ma from ±5V to ±15V supplies. Input and output offset voltages are kept to within 10 mV of ground reference.

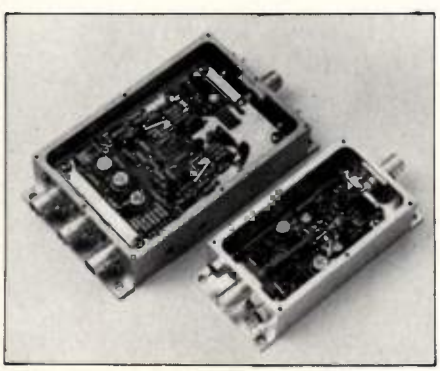
Also available is the CLC101 which features true differential inputs and a DC to 150 MHz bandwidth. Passband flatness from DC to 100 MHz is ±.5 dB max, while delivering +18 dBm into 50 ohms. THD is less than .3 percent at 10 kHz, and less than 1 percent up to 20 MHz at +24 dBm into 50 ohms or at 20 Vpp into an open circuit. As a pulse amplifier, the output can slew at 3300 V/μs and deliver 16 Vpp into 50 ohms. For a 10V step, rise time is 3 ns, overshoot less than 10 percent for a duration of less than 10 ns, and settling time to 1 percent is less than 1 μs. Nominal gain is 29.6 dB, input and output impedances are 50 ohms, and supply current is 190 ma from ±15V to ±24V supplies.

Special features of the CLC101 are a DC offset adjustment which can be used to offset the output waveform by as much as ±5 VDC, and operate as a limiter. When the inputs are overdriven, a limited version of the input waveform results. An adjustment is provided to vary the amplitude of the output over a 20 dB range. The use of these features allows the CLC101 to drive any logic family from ECL to TTL to CMOS at logic levels determined by the user.

Both the CLC100 and the CLC101 contain internal voltage regulators so that well regulated power supplies are unnecessary. Also, inputs and outputs are well protected against short circuits and overload conditions.

The unique performance capability of these amplifiers is made possible by combining a high frequency fixed gain amplifier in such a way that they work to complement each other. One other beneficial result is that the high gain low frequency amplifier stabilizes the DC offset and distortion of the high frequency amplifier so that temperature changes have little effect.

Contact Comlinear Corporation, 514 Railroad Ave., Loveland, CO 80537. (303) 669-9433. INFO/CARD #110. ☐



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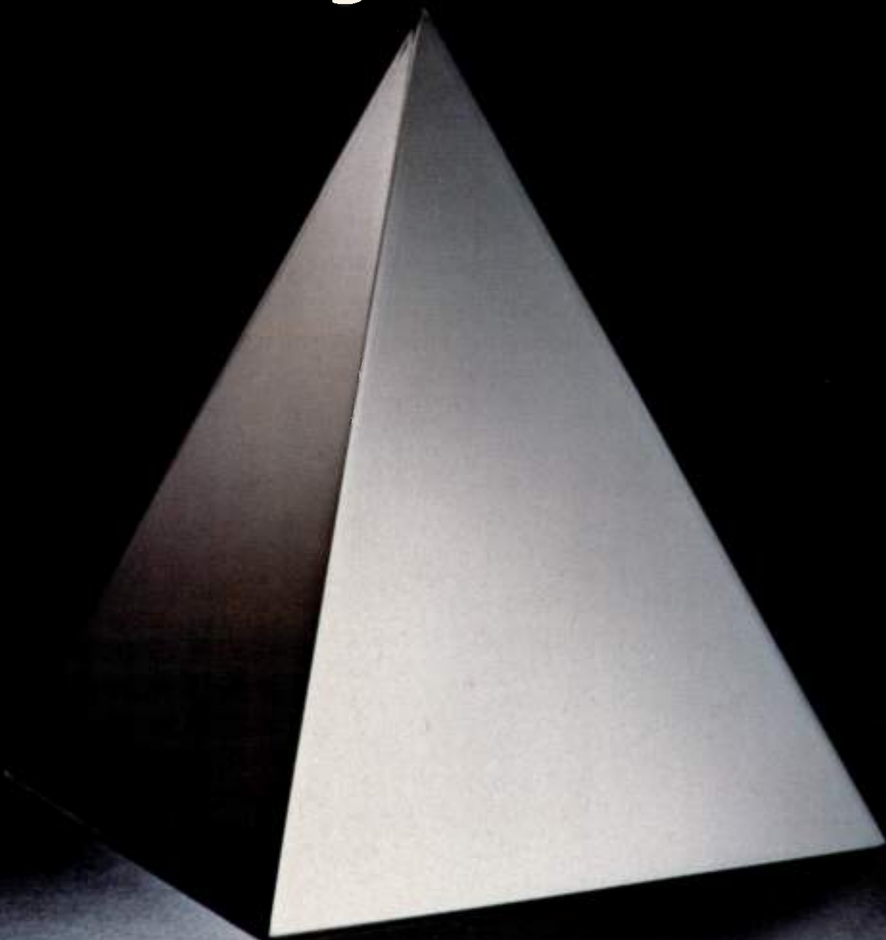

The new ENI 550L delivers 50W, 1.5-400 MHz.

This single unit is so incredibly versatile it can replace several you may be using now. And you may never need another. It's an extremely broadband high power, solid state, Class A linear amplifier. It's rated at 50W from 1.5-400 MHz. But it can provide 100 Watts from 1.5-220 MHz. All you need with the 550L is any standard signal or sweep generator and you've got the ultimate in linear power for such applications as RFI/EMI testing, NMR, RF Transmission, ultrasonics and more.

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The 550L represents the pinnacle in RF power versatility. There's nothing like it commercially available anywhere! And it may be the only RF power amplifier you ever need.

For more information, a demonstration, or a full line catalog, please contact us at ENI, 3000 Winton Road South, Rochester, NY 14623. Call 716/473-6900, or telex 97-8283 ENI ROC.

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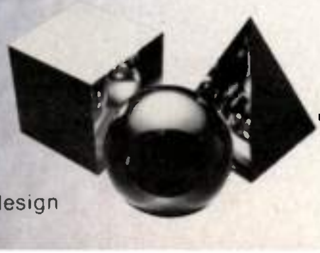

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