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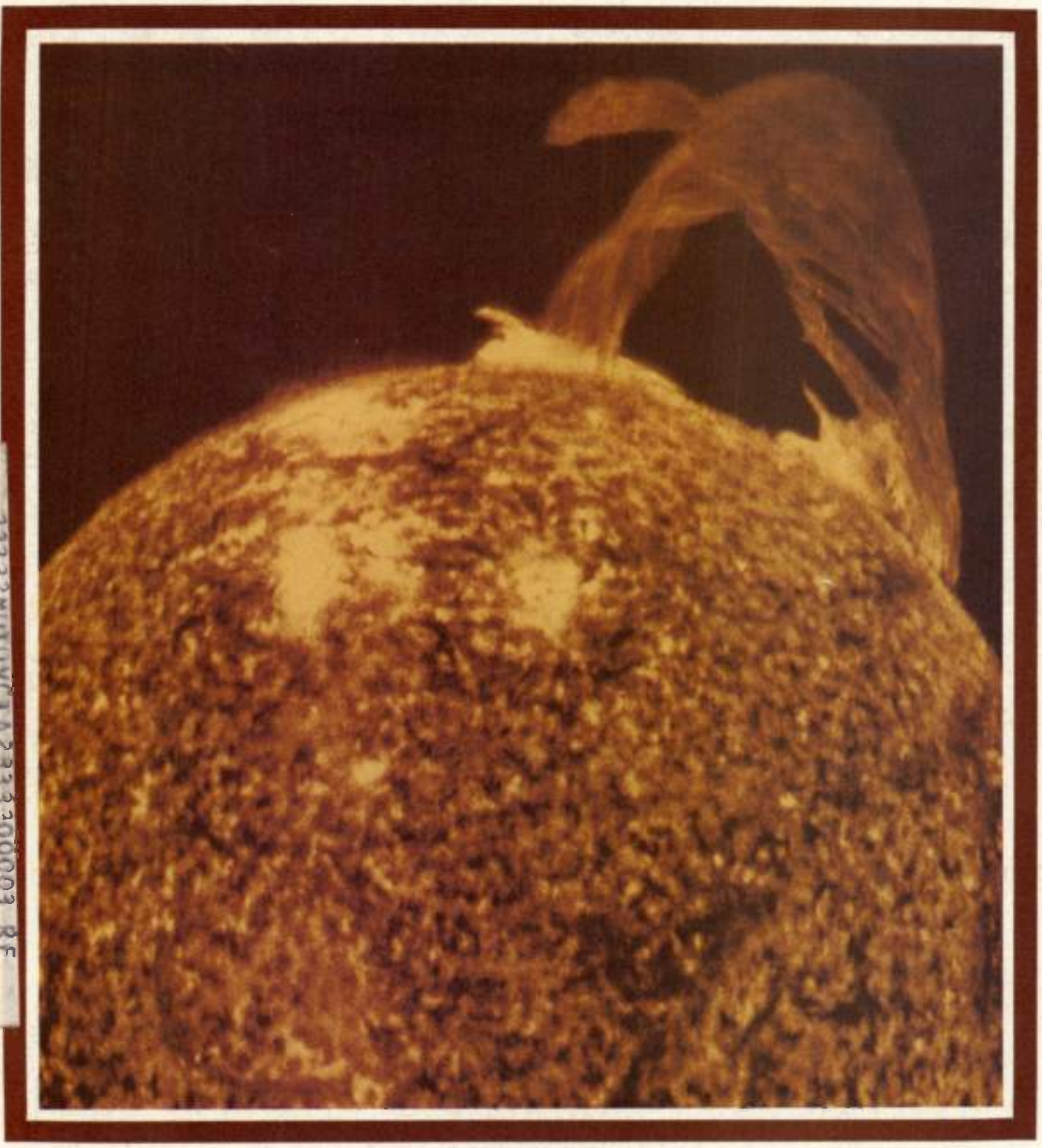

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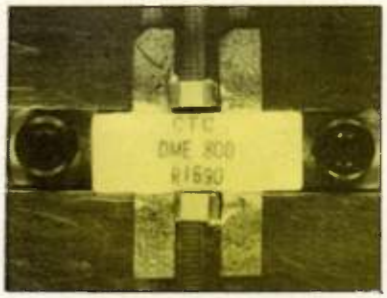
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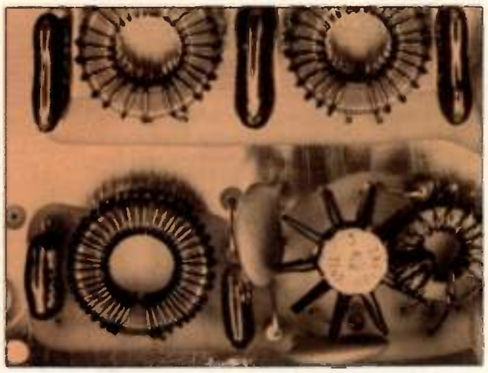
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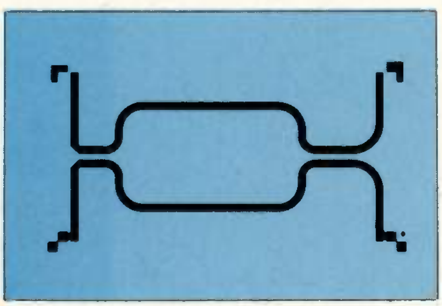
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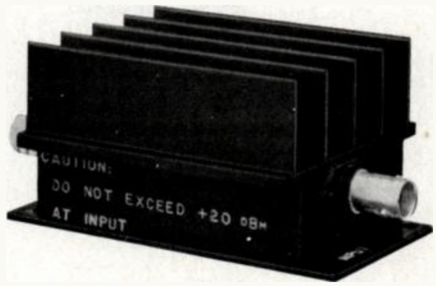

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Rich Rosen

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Patty Heil
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The Maunder Minimum Or What is Constant?

Recently, I had an enjoyable conversation with an interesting and intelligent staff scientist at one of the several solar observatories in the world. The conversation, though centered on our solar system "the sun", ran the gamut from the basics of the scientific method to Einstein's Special Theory of Relativity in 1905. It *even* touched on religion. Among other things, I learned that in their daily work, many phenomena that they observe fall under the category of "unexplained" (with shades of "twilight zone" music playing in the background); that there are "fruit" stars — so labelled because of their "fruity" or highly unusual type of behavior.

In this last respect, our sun, the one we depend upon so very heavily for its multi-faceted energy output, though considered relatively constant in the microscopic time point of view, has very recently modified its behavior. It appears that for millenia it has regularly behaved in an *irregular* fashion. The apparent periodicity, so prevalent in the 11-year cycle of solar activity — with its pronounced effect on global communications, just did *not* always exist. A technique related to Carbon 14 datings, in tree rings, indicates that for many thousands of years our sun generated energy and had magnetic sunspot events at an apparent irregular rate. Suddenly in the 17th century it "cranked down" — producing a lull in activity or what some people have referred to as a mini ice age. The number of visible sunspots in a 70 year period from 1645 to 1715, known as the Maunder Minimum, were negligible. Then, just as suddenly the sun started acting regularly, with its present day 11 year cycle as an outstanding feature.

Consider that it takes you 1 hour to skim through this magazine. For "59 minutes and 59 seconds" the sun has acted one way. In the very last "second," since your great-great grandfather's time till now, *it's changed its behavior.* To the best of my knowledge and experts in the field — *it can't be explained.*

Man does not live by bread alone and the intellectual enjoyment felt in discussing and pursuing some of nature's *constants* is almost indescribable.

Speaking of bread I so very much enjoyed calling up Mr. Len Roland of Tektronix the other day and informing him that he won the $1000 savings bond in the January ad survey contest. (More about Len and the results of the contest on page 17.)

Electro/80 is coming, May 13-15 in Boston this year. Please look for me (see photo) and you will probably find me at the RF technical talks (reviewed in this issue) or any of the RF exhibitors' booths. See you then.

Rich Rosen

Rich Rosen

Thermal Resistances Related To Flange Package Mounting

Methods and merits of heat removal from RF high power transistors are discussed.

Howard Bartlow
CTC, San Carlos, Calif.

The problem of heat removal from high power RF transistors has been around since the beginning of their use in amplifiers. Various methods of getting heat from package to ambient air have been devised including radial fins on metal cans, stud packages and flange packages. Typically for highest power dissipation, flange packages have the advantage; and it is to this type of flange-heat sink system that this report is addressed.

In an attempt to quantify various qualitative judgments of what is and is not important; as well as what surfaces are best for flange heat sinking, test equipment was designed and a simple test developed. This involved starting with a machined perfectly flat aluminum heat sink, with threaded holes mated to the flange hole spacing. In this heat sink a .039 inch hole was drilled very close to the outer perimeter contact point of the package to a depth of .10 inches. An identical hole was drilled in the circular portion of the flange (Figure 1). Two matched thermocouples which were calibrated to read identically for equal temperatures were buried in thermal compound (GE Insulgrease G-641 Silicone Dielectric Compound) and inserted in the flange and heat sink holes. A switch was installed between the thermocouples and the temperature meter. It was then possible to read case and heat sink temperatures with very good relative accuracy and with minimal time lapses between readings by switching from case to heat sink and back to case. Readings were taken at 20-30 second intervals for periods up to five (5) minutes. Flange to heat sink temperature differences stabilized in approximately twenty (20) seconds and higher.

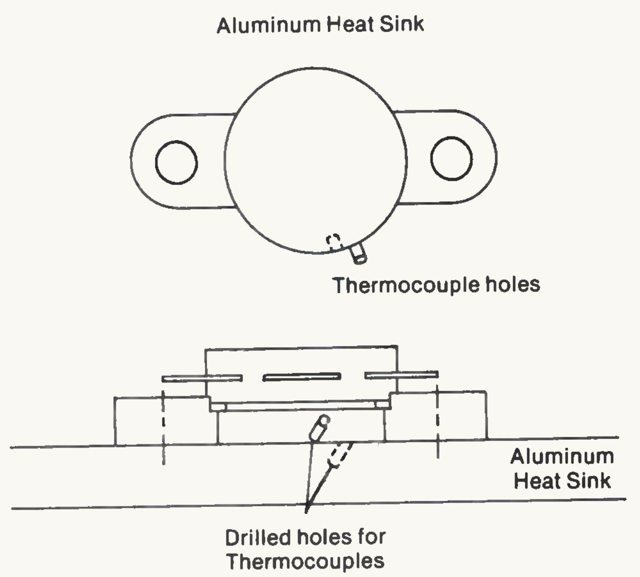

Figure 1. Temperature Measuring Technique.

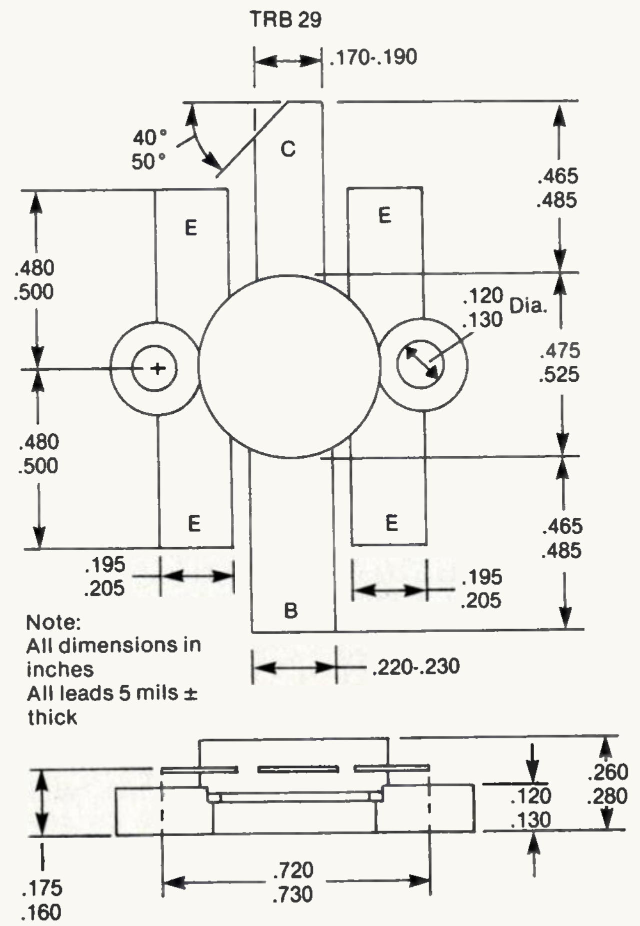

Figure 2. UHF Transistor Utilized.

Several experiments have been run using a popular UHF input matched transistor package of these dimensions (Figure 2). Eighty watts was dissipated by the device through the flange to heat sink. No attempt was made to maintain heat sink temperatures at a constant level since the delta (temperature difference flange-to-heat sink, Δ) was the only measured parameter of importance and since water or forced air cooling can easily introduce gradients and consequent errors. Additionally, it was found that this delta was consistent through heat sink temperature excursions of 25°C to 100°C.

One accuracy factor emerged after some testing — it is important that all thermocouple holes be nearly equal in depth. Other normal precautions were observed and test repeatability appears to be quite good.

Flange flatness is measured using a calibrated contact gauge system which indicates maximum variation from nominal (flat). This is done in several passes in both X and Y directions. The result is recorded as **total indicated reading** (t.i.r.) and increases as flange flatness decreases, (i.e. higher t.i.r. packages are less flat then lower t.i.r. packages). Worst case measurements were done with packages reading ≈ 1.8 mil t.i.r. Best case readings (including sanded flanges) were ≈ .3 mil t.i.r.

Experiments Performed — Thermal Compound

First and foremost, it has been found that a good thermal compound always improves the $\Theta_{f\text{-}hs}$. Also, the improvement is greater for packages having higher t.i.r. number (≈1.5 mil). Quantitatively, very flat packages (t.i.r.≈ .2 mil) were improved approximately .1°C/W by using thermal compound while worst case packages (t.i.r.≈1.7 mil) were improved nearly .23°C/W over these same packages with no thermal compound. Therefore, the effects of package flatness on $\Theta_{f\text{-}hs}$ are reduced when thermal compound is used.

No special precautions (other than cleanliness, freedom from grit etc.) are necessary when using thermal compound. Further, there was no evidence whatsoever of cavity "packing" of the compound holding the flange away from the heat sink and consequent increases in $\Theta_{f\text{-}hs}$.

Rather, in every case significant reductions in $\Theta_{f\text{-}hs}$ were noted by using thermal compound. This is probably because the compound fills in microscopic pores and irregularities of flange and heat sink thereby increasing heat transfer surface area.

Torquing Method

Flange heat sinking benefits from a proper torquing method in that certain compression and torsional stresses combined on copper flanges often produce significant flatness changes. These stresses can occur whenever surface irregularities, tapers, ridges, etc. are present — as they may be on both bolts and flanges. To counteract this effect, bolts should be tightened similar to "lug" bolts on automobile wheels. The bolts should be snug tightened (at least finger tight) before torquing to spec.

If one bolt is dropped in and torqued to specifications; then the second is dropped in and torqued also, an *increase* in $\Theta_{f\text{-}hs}$ of up to .27°C/W will be observed. The correct procedure is to snug (even if only finger tight) both bolts then torque both.

Torque Specification

Using 4 in-lbs., 6 in-lbs., and 8 in-lbs. of torque (and a few tests to 10-12 in-lbs.), several comparison tests were run. Virtually no differences were observed. However, on certain packages a very slight improvement trend resulted when increasing from 4 in-lbs. to 6 in-lbs. Therefore, we recommend 6 in-lbs. and are confident there is considerable tolerance in this specification.

Sanded Flanges and Oxidized Copper

Careful hand sanding or lapping will typically bring all packages to approximate t.i.r.'s of ⩽.3 mil. This method is sometimes employed in the industry at present.

In our tests, it did indeed tend to bring all tested samples within .04°C/W of each other in $\Theta_{f\text{-}hs}$.

The effect of copper oxide on this sanded surface was also measured by purposely oxidizing the flange at high temperatures (>200°C) until an obvious layer(s) was formed. The device was retested; then resanded, and tested again. No effect on $\Theta_{f\text{-}hs}$ was found even with this rather severe treatment.

The test results indicate it is not necessary to take special precautions to prevent copper oxidation of sanded flanges prior to use.

$\Theta_{f\text{-}hs}$ Differences in Typical Packages

Special selection of worst case, typical and best case t.i.r. packages in three separate trial comparisons has enabled a prediction of probable difference in $\Theta_{f\text{-}hs}$ for each type. Samples of .2 mil, .7 mil and 1.7 mil t.i.r. packages were used and the following deltas were recorded.

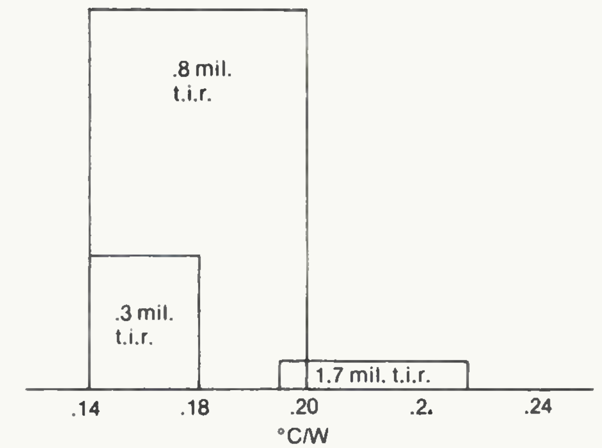

Figure 3. $\Theta_{f\text{-}hs}$ Differences from Flange Flatness.

The maximum difference between .2 mil and 1.7 mil packages was a surprisingly low .09°C/W (using thermal compound). The typical .7 mil t.i.r. packages read within .06°C/W of the best case measurement and in some samples equaled the sanded flat (<.3 mil t.i.r.) packages. Absolute numbers ranged from approximately .14°C/W (.3 mil t.i.r. package) to .23°C/W for the worst case 1.7 mil t.i.r. package (Figure 3).

It appears that a t.i.r. specification for flatness of 1 mil or less is quite adequate.

Accuracy of $\Theta_{f\text{-}hs}$ Measurements

By checking repeatability and testing a significant quantity of packages, it is estimated that $\Theta_{f\text{-}hs}$ absolute accuracy is ⩽.03°C/W and relative (delta) accuracy is ⩽.02°C/W in these test measurements.

Conclusions — Recommendations

For the most consistent results and lowest $\Theta_{f\text{-}hs}$ measurements, the following recommendations should be followed when mounting flange packages:

1. Always use thermal compound taking care to eliminate any grit or dirt on surfaces.
2. The torque method should always be to snug (at least finger tight) both bolts, then torque both.
3. A torque specification of 6 in-lbs. is recommended.
4. Though not usually necessary sanding of non-flat packages is acceptable. Special precautions to prevent copper oxidation do not appear necessary.
5. A 1 mil t.i.r. specification for flatness is sufficient and recommended. □

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Electro/80

Tuesday, May 13 is an outstanding day for RF Engineers taking part in the diversified 68 hour professional program in Boston this year.

Electro/80 held in Boston this year, starting May 13th, has an electronics show with 482 exhibitors (as of Jan. 17th), a professional program featuring 34 technical two hour long sessions, seminars, workshops, luncheons and a film theater.

According to the preliminary technical program schedule Electro/80 has something of interest for everyone. The *r.f. design* editor, yours truly, is looking forward to sessions, 2, 5 and 8 on Tuesday. Each session's content is detailed below.

Tuesday, May 13th

9 AM-11AM

Session 2 Integrated Solid-State Microwave Amplifier Design

2/1 Microwave properties of materials for M.I.C.
2/2 Basic design parameters for microstrip circuits
2/3 Low noise GaAs FET amplifier design
2/4 High power Impatt amplifier design and applications
2/5 Monolithic circuit design

Noon-2 PM

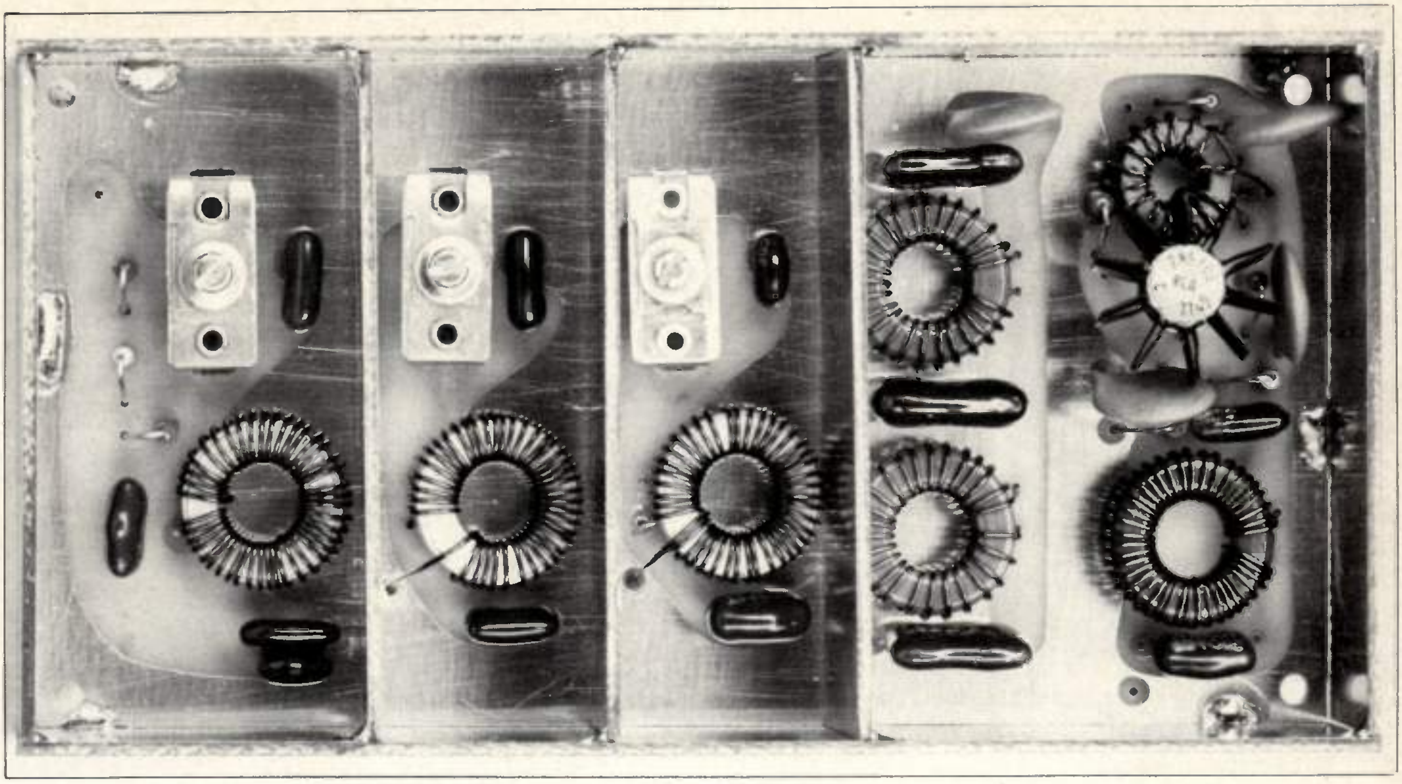
Session 5 Recent Developments in Communications Receiver Design

5/1 Recent developments in communications receiver design to increase dynamic range.
5/2 Large and small signal applications for FETs in modern communications receivers.
5/3 The real world of high performance receivers.
5/4 Applying VMOS power FETs in high level balanced mixers.

3 PM-5 PM

Session 8 Current Developments and Applications in the RF Power Device Field.

8/1 Transmission line transformers with low transformation ratios
8/2 The present state of the VMOS power FET art.
8/3 Parasitic oscillation in solid-state RF power amplifiers — causes and cures-demystified.



Magnetic Cores in RF Circuits

Understanding core parameters helps the RF Engineer in designing inductors and transformers.

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

Getting the proper ferrite or powdered-iron toroid, sleeve or pot-core into your RF circuit is not as casual an exercise as some designers believe! We must consider the core performance versus operating frequency (Q_L and bandwidth), the power-handling ability of the chosen core and the physical size of the completed inductor or transformer. A factor of equal importance in many circuits is the core-material stability in the presence of heat (ambient and current-induced heating). The design approach has been confusing to some RF engineers and technicians who have not had the occasion to develop circuits in which magnetic-core devices were required. This paper will, therefore, address the matter of fundamental design procedures for narrow and broadband RF inductors and transformers which contain magnetic core materials. For the most part, the discussion deals with the practical aspects of the design exercise.

Powdered Iron or Ferrite?

A notable advantage of powdered-iron over ferrite-core material is that the powdered iron B_{op} in gauss (operational flux density) is much greater for a given cross-sectional area than that of ferrite. Owing to this virtue the size of an inductor or transformer can be substantially smaller if iron is used rather than ferrite. In essence, powdered-iron cores yield higher values of B_{max} (maximum flux density in gauss) than the ferrite cores of equivalent physical dimensions.

Another advantage in the use of powdered-iron cores is *core stability*. Generally, ferrite materials are more prone to permeability changes with temperature than is the case with iron. Furthermore, if the B_{max} of a ferrite core is exceeded in an RF power circuit the permeability of the material will undergo a permanent change. This is not true of powdered-iron core material.

Why then should the designer consider using ferrite core material when iron seems to be so much better? Paramount among the decisions to employ ferrite cores is the very high value of μ_i (initial permeability) which is available: Ferrite cores can be purchased with permeability factors in excess of 5000. The common upper limit for powdered-iron cores is approximately 75 in terms of μ_i. In view of the foregoing it becomes apparent that ferrite materials lend themselves well to the fabrication of low and medium-frequency inductors and transformers. This is especially significant with respect to broadband RF transformers. The principle is illustrated dramatically in

Figure 1 where two inductors are wound for use at 455 kHz on 0.5-inch OD toroid cores, using powdered iron and ferrite materials for the frequency of operation versus good Q. A substantially greater number of turns is needed for L1 than is required for L2. The advantage of the ferrite over the powdered iron is clear: In the case of L2 we can use fewer turns of heavier gauge wire. The net result is higher tuned-circuit Q and reduced fabrication time. Both inductors have been calculated for single-layer coil windings.

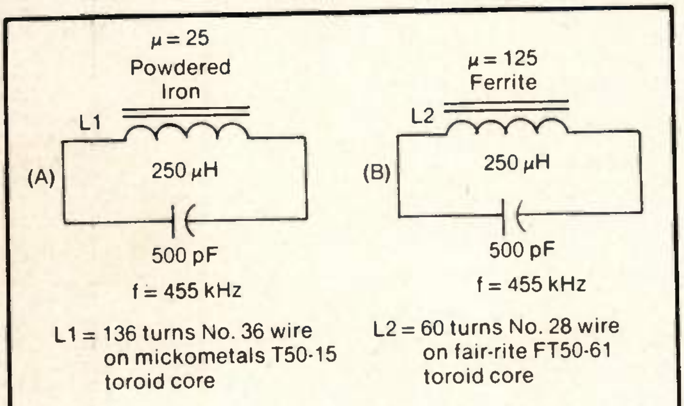


Figure 1.

Powdered-iron core materials have been characterized mainly as being best suited to applications at medium frequency and higher. Even though the irons have been used successfully for many years as low as a few hundred Hertz, their greatest application has been at RF. This is because it is possible to obtain low values of μ_i with powdered iron (as low as 1, for use at VHF). This permits the designer to obtain suitable values of Q_u through the VHF spectrum, which is not possible with ferrite core materials. For example, the Micrometals 0-mix iron is recommended for use between 50 and 300 MHz. It has a μ_i of 1. The next frequency bracket (20 to 250 MHz) lists 12-mix as the optimum core material. It has a μ_i of 3. It can be seen from this that the higher the μ_i the lower the recommended operating frequency in terms of maximum attainable Q_u. At the opposite end of the scale we find the Micrometals mix-41 iron, which has a μ_i of 75. The suggested frequency range for this material is 1 to 10 kHz.

Fair Rite Corporation produces a material-68 ferrite core which has a μ_i of 20 and is specified for use between 80 and 180 MHz. Most designers seem to limit their use of ferrites to a maximum frequency of approximately 25 MHz when designing narrow-band circuits. A suitable ferrite material for this range is mix 63. It has a μ_i of 40. The manufacturer suggests using mix 63 from 15-25 MHz.

Why Use Magnetic Cores at RF?

There are a number of advantages we can realize by using magnetic core material in RF inductors and transformers. When we consider the alternative — air-wound solenoidal inductors — the advantage becomes obvious. The magnetic-core component allows the mass of the circuit to be reduced markedly.

Still another advantage to be realized is seen in the case of toroidal inductors. As opposed to conventional inductors they are *self-shielding*. This eliminates the need for metal shield enclosures around the coil or transformer. The same is true of pot-core inductors and transformers. This advantage is lost when rods are used as the core material.

A more subtle advantage attendant to magnetic cores is improved Q_u when large amounts of inductance are required. This is because an equivalent air-core inductance would require many turns of light-gauge wire, whereas the magnetic-core equivalent inductance could be realized with substantially fewer turns of heavier-gauge wire. This would reduce the series R component to enhance the Q.

Broadband Transformers

Perhaps the greatest use of magnetic cores in RF circuits is found in broadband-transformer applications. These take the form of *conventional* and *transmission-line transformers*. They are commonly used in solid-state transmitters and receivers. They also lend themselves to viable use in antenna systems as baluns, combiners and matching transformers. Toroids, balun cores and rods are suitable for use in antenna systems. Pot cores are used as well as the foregoing for small-signal broadband transformers.

Figure 2 illustrates the basic difference in the winding format of conventional and transmission-line transformers. Illustration A shows a conventional transformer schemat-

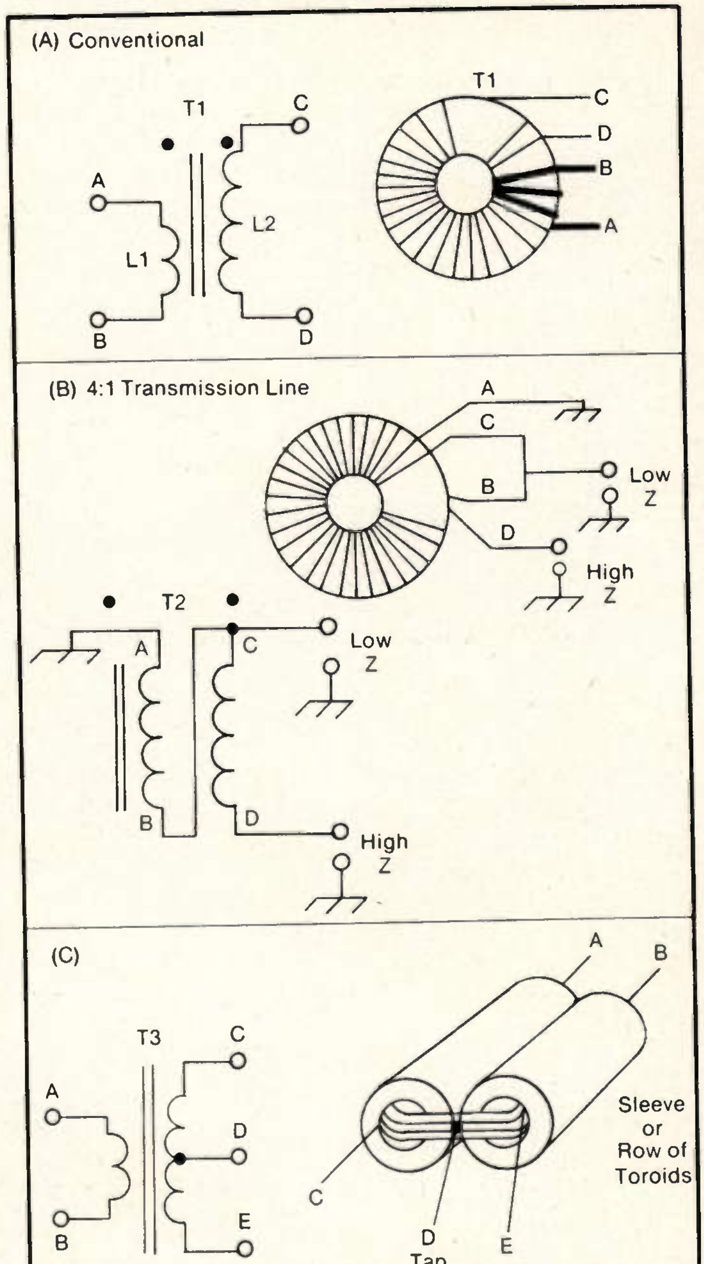


Figure 2.

ically and pictorially. The major advantage of this style of broadband transformer is that a wide latitude of transformation ratios are possible, since the primary and secondary windings are separate. The consensus is that conventional transformers are not as efficient as transmission-line equivalents are. Figure 2C illustrates another common form of conventional transformer. It uses two rows of toroid cores (stacked) or long ferrite sleeves (2 each). This type of broadband transformer is used frequently in solid-state RF power amplifiers for input and output matching.

A simple illustration of the transmission-line transformer is seen at B of Figure 2. The example shows an unbalanced-to-unbalanced 4:1 transformer which contains a bifilar winding. The wires can be laid on the core in parallel or they can be twisted an appropriate number of times per inch. Alternatively, lengths of subminiature coaxial cable are sometimes used. Microdot 25-ohm line is frequently the designer's choice. The disadvantage in using this style of transformer is that only fixed transformation ratios are possible — 1:1, 4:1, 9:1, 16:1, etc. It is recommended that a winding gap of approximately 30° be left on the toroid core as shown in Figure 2 at A and B. This technique aids in reduction of unwanted distributed capacitance across the winding and between the coil turns. The benefits from maintaining a gap in the winding are most pronounced when large numbers of turns are laid on a core. For example, a 900-turn winding might exhibit 70 pF of distributed capacitance with the coil ends directly adjacent to one another. The same inductor, but with a 30° winding gap, might show only 40 pF of distributed capacitance. The undesirable effects of large distributed capacitances are high "effective inductance" and degraded Q. In this context, the closer the coil's self-resonance is to the operating frequency, the poorer the Q.

Winding Characteristics

We must ensure that the individual windings of a broadband transformer exhibit sufficient reactance (X_L) at the lowest operating frequency to be effective as transformers or chokes. A viable rule of thumb is to make the smallest transformer winding approximately 4 to 5 times higher in X_L than the load resistance connected to the winding. Hence, if we were to match a 50-ohm source to a 10-ohm load with a 5:1 broadband transformer, the 10-ohm winding should have an X_L of approximately 40 to 50 ohms at the lowest operating frequency. This is demonstrated in Figure 3. L2 of T1 looks into a 10-ohm load at the base of Q1. Using the X4 rule, the winding reactance must be 40 ohms. In order to ensure ample winding inductance

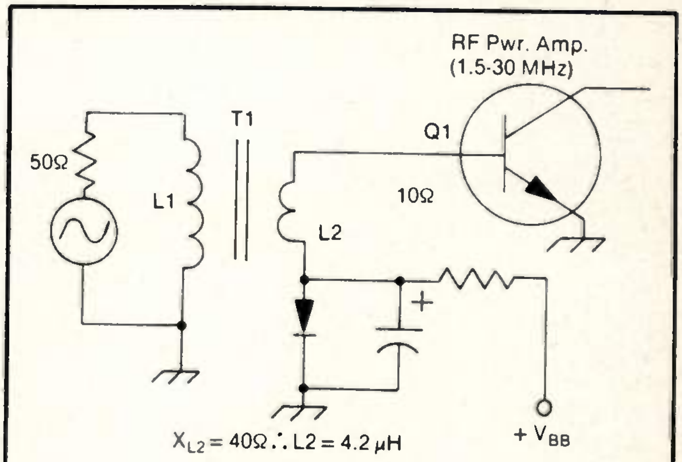


Figure 3.

at 1.5 MHz we will need 4.2 μH. Once this requirement is satisfied, L1 will also have sufficient X_L for the 50-ohm source impedance of the generator. Too many transformer turns could lead to excessive X_c, as discussed earlier. This might result if we followed the X10 X_L rule that has been suggested by some designers.

Ferrite core material is used almost exclusively in broadband transformers. Generally, the materials are of fairly high μ_i, such as 950 for use between 1.5 and 30 MHz. This is done to ensure ample inductance in the windings at the lower end of the spectrum being covered. As the operating frequency is elevated the core effects begin to diminish, thereby giving rise to excellent transformer characteristics at the upper end of the desired frequency range. The actual permeability chosen will depend largely upon the frequency spread desired and the operating-frequency limits.

The Proper Size Core for the Job

The undesirable effects of using too small a core are manifested in several ways. First, saturation will take place at or below peak-power level. Examination of the hysterisis profile of a saturating core will reveal a perfect condition for the enhancement of harmonic currents. The output waveform of an RF transformer which is saturating will show a square-wave characteristic when viewed on a scope. This is definitely an unwanted characteristic in most RF circuits.

Saturation changes the μ_e (effective permeability), which can lead to instability and losses. In a worst-case situation the core can become permanently damaged or might even shatter. Ideally, the core should be operated in the linear portion of its B_{max} curve. This data is available from the manufacturer.

The first consideration in determining the required core size is the B_{op} that will prevail. This can be found from

$$B_{op(ac)} = \frac{E_{rms} \times 10^8}{4.44\, f\, N\, A_e} \text{ gauss} \qquad (1)$$

where B_{op} is the operational flux density, E_{rms} is the voltage across the winding, N is the number of turns, A_e is the equivalent area of the magnetic path in cm^2 and f is in Hz.

The situation changes somewhat when both AC and DC currents flow in the winding. $B_{op(total)}$ is obtained from

$$B_{op(total)} = \frac{E_{rms} \times 10^8}{4.44\, f\, n\, A_e} + \frac{N\, I_{DC}\, A_L}{10\, A_e} \text{ gauss} \qquad (2)$$

where I_{DC} is the direct current through the winding and A_L is the manufacturer's inductance index for the specified core.

This requires some knowledge of the operating parameters which will prevail before a trial core is chosen for the equations. As cumbersome an approach as this is, it becomes part of the design procedure. If a given core will not yield the required safe margin between B_{op} and B_{max}, then the next larger size core is selected for the equations.

Some designers prefer to use E_{pk} in equations 1 and 2. This allows a safety factor beyond that which the rms value of the voltage will allow. However, the writer prefers to use the largest core that will fit into the available space on a circuit board or equipment module. This usually permits the designer to constrict B_{op} to the linear portion of the core's curve. The price differential between an electrically suitable core and the next larger size is minimal.

The Correct Number of Turns

Regardless of the physical structure of the magnetic core material being used, each core type and size has a specific A_L factor which can be used to determine how many turns are needed for a given inductance. A_L information is published by most manufacturers of magnetic core material. If the A_L factor for a given core can't be obtained, it can be determined easily by placing a few trial turns of wire on the unknown core and measuring the inductance of the winding. A_L can then be obtained from

$$A_L = L_{\mu H} \times 10^4 / N^2 \tag{3}$$

where $L_{\mu H}$ is the measured inductance and N is the number of turns. Thus, if we wound 6 turns of wire on an unknown toroid core and the measured inductance was 1.0 μH, the A_L would equal

$$1.0 \times 10^4/36 = 277.7$$

Once the A_L factor is known it is an easy matter to calculate the turns necessary for a given inductance. For example, assume that we used the core for which we obtained the A_L factor by means of Equation 3. We desire an inductance of, say, 112 μH. The winding size can be determined from

$$N = 100\sqrt{L_{\mu H}/A_L}\ turns$$

hence

$$N = 100\sqrt{112/277.7} = 63.5\ turns \tag{4}$$

When dealing with large amounts of inductance on cores with relatively high permeability factors, it is common to change Equation 4 to

$$N = 3162\sqrt{L_{mH}/A_L}\ turns \tag{5}$$

These equations apply to any magnetic core material — toroids, pot cores and rods.

Summary

Additional applications information and design data can be obtained from the publications listed in the bibliography. Also included are the names and addresses of some leading U.S. manufacturers of ferrite and powdered-iron core materials. Their literature contains detailed information which will be of use to the designer. □

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Hayward and DeMaw, *Solid-State Design for the Radio Amateur*. ARRL, Inc., Newington, CT 06111.

Hilberts, *Amperex High-Power Transistor Book*, "On the Design of HF Wideband Power Transformers," *ECOM* report 6907.

Polydoroff, *High-Frequency Magnetic Materials*, John Wiley and Sons, N.Y.

Practical Circuit Information:

The ARRL Radio Amateur's Handbook, 1980 Edition, ARRL, Inc. Newington, CT 06111. Contains numerous applications of magnetic cores in HF and VHF transmitting and receiving equipment.

Manufacturers and Distributors:

Amidon Associates, 12033 Otsego Street, N. Hollywood, CA 91607. Single lot sales by mail plus catalog and data sheets.

Arnold Engineering Co., Box G, Morengo, IL 60152 (mfgr.)

Elna Ferrite Labs, Inc., 9 Pine Grove, Woodstock, NY 12498 (small-quantity, by-mail distributor).

Fair-Rite Products Corp., 1 Commercial Row, Walkill, NY 12589 (mfgr.)

Ferroxcube Corp., 5083 Kings Hwy., Saugerties, NY 12477 (mfgr.).

Indiana General Ferrite Products, Crows Mill Rd., Keasby, NJ 08832 (mfgr.).

Magnetics, Div. of Spang Industries, Box 391, Butler, PA 16001 (mfgr.)

Micrometals, Inc., 1190 N. Hawk Circle, Anaheim, CA 92807 (mfgr.).

Palomar Engineers, Box 455, Escondido, CA 92025. (small-quantity, by-mail distributor).

Congratulations Len

Meet Len Roland, 37 years old, a Microelectronics component design engineer with Tektronix. He has a BS in Math and EE. Spent four years in the army in Fairbanks, Alaska repairing radio teletype equipment, taught Math at Portland State University and a broad electronics curriculum at United Electronic Institute. Presently he is designing thick and thin film hybrid, broadband amplifiers and drivers.

We are just as glad Len took the time out to judge the top eight ads as he is to win the savings bond.

This is how we determined the top eight ads:

928 Engineers responded to the January **r.f.design** ad survey contest. All their cards formed the basis in determining the order of the top eight ads (out of twenty for that issue) and consequently the winner Len Roland. The method utilized can best be described as a cumulative weighting. Here's how it worked: The ad "A Spectrum Analyzer With . . .", by Tektronix, received the following votes, see Table 1.

Table 1 — Cumulative Weighting

Place	Votes	Weighting	Subtotal
1	318	8	2544
2	257	7	1799
3	127	6	762
4	89	5	445
5	58	4	232
6	25	3	75
7	7	2	14
8	16	1	16
			5887 Total

5887 is the highest cumulative weighted total and consequently determined the number one ad as Tektronix's. Using this method (there are others, we realize) the top eight ads are listed below in Table 2.

Table 2. Top Eight Ads

Position	Company	Ad
1	Tektronix	"A Spectrum Analyzer With . . ."
2	Communication Specialists	"A Fresh Idea . . ."
3	John Fluke Mfg. Co.	"Engineering designed this new counter . . ."
4	Microwave Associates	"Don't let the atmosphere . . ."
5	Texscan	"Texscan sets the world standard . . ."
6	Optoelectronics	"When Quality Counts . . ."
7	Solid State Microwave	"When it comes to high reliability . . ."
8	Hewlett-Packard	"For high performance receiver . . ."

Len Roland and several others all agreed on the first four ads in the proper sequence. Since no one picked the number five spot correctly the cumulative weighting total method was again utilized to determine the highest score for the last four ads.

Once again congratulations Len.

Tektronix 492 Spectrum Analyzer

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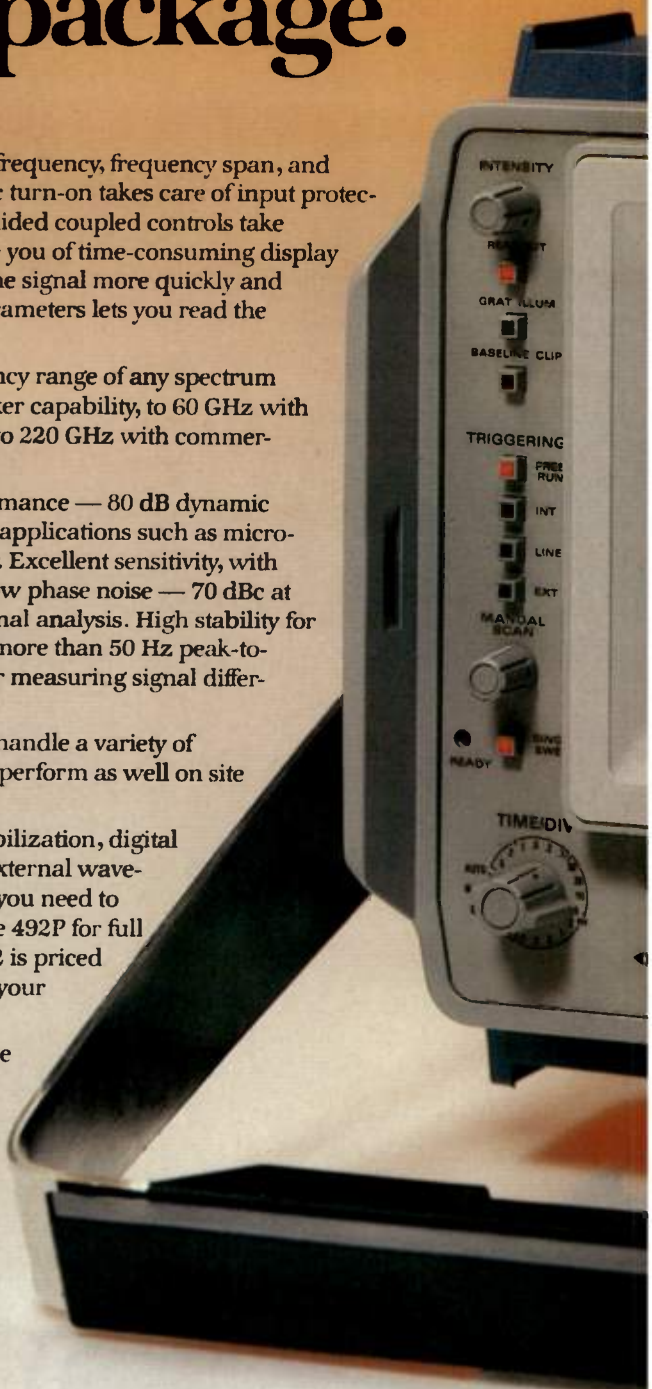

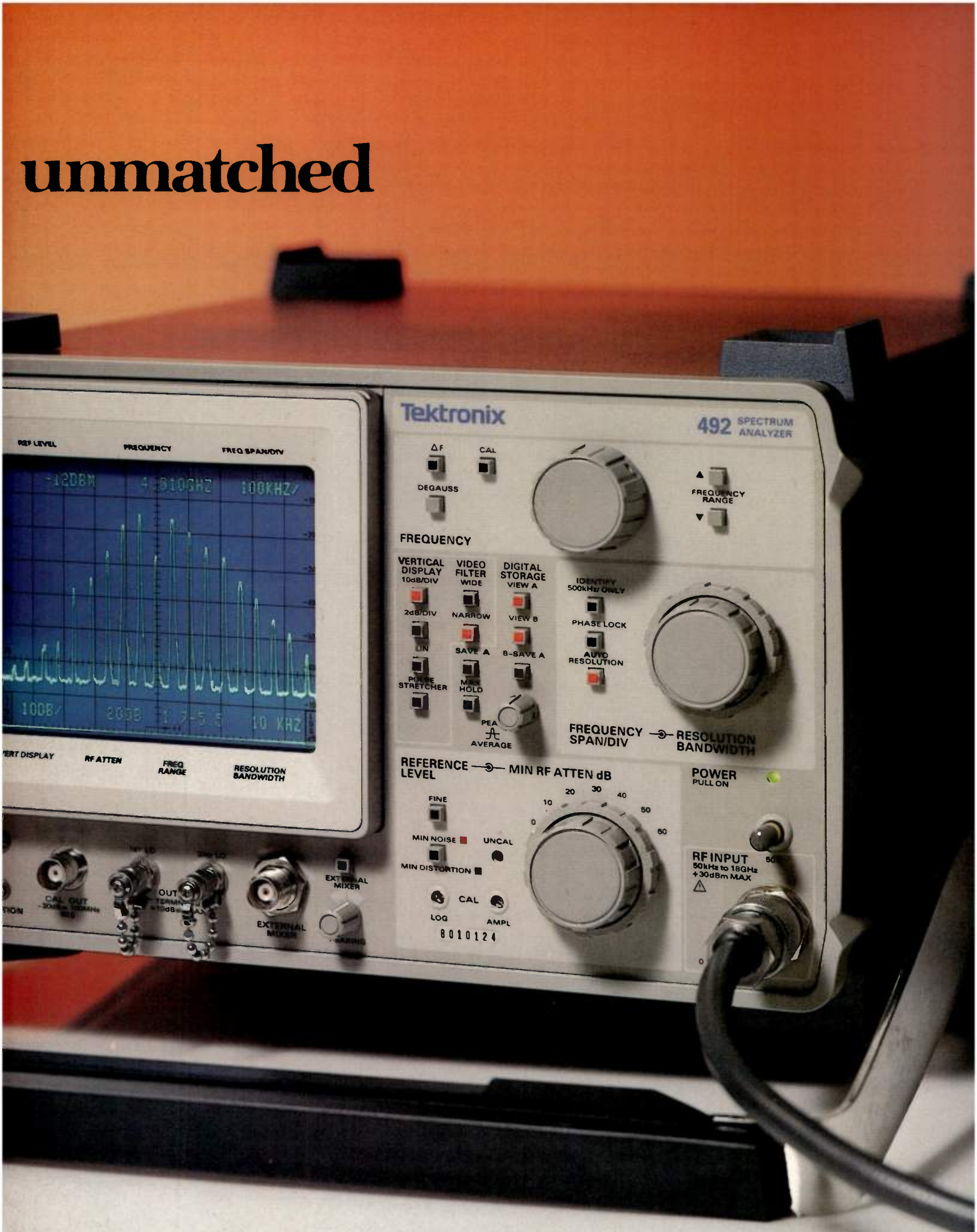
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8010 *8010.1	325.00 405.00	1 GHz	9	1-10 mV	5-20 mV	10-25 mV	1-10 mV	[8] .01-20 SEC	.1 Hz	1 Hz	10 Hz 1 GHz	1 PPM 0.1 PPM	10 MHz	YES STD	YES OPTION $39.

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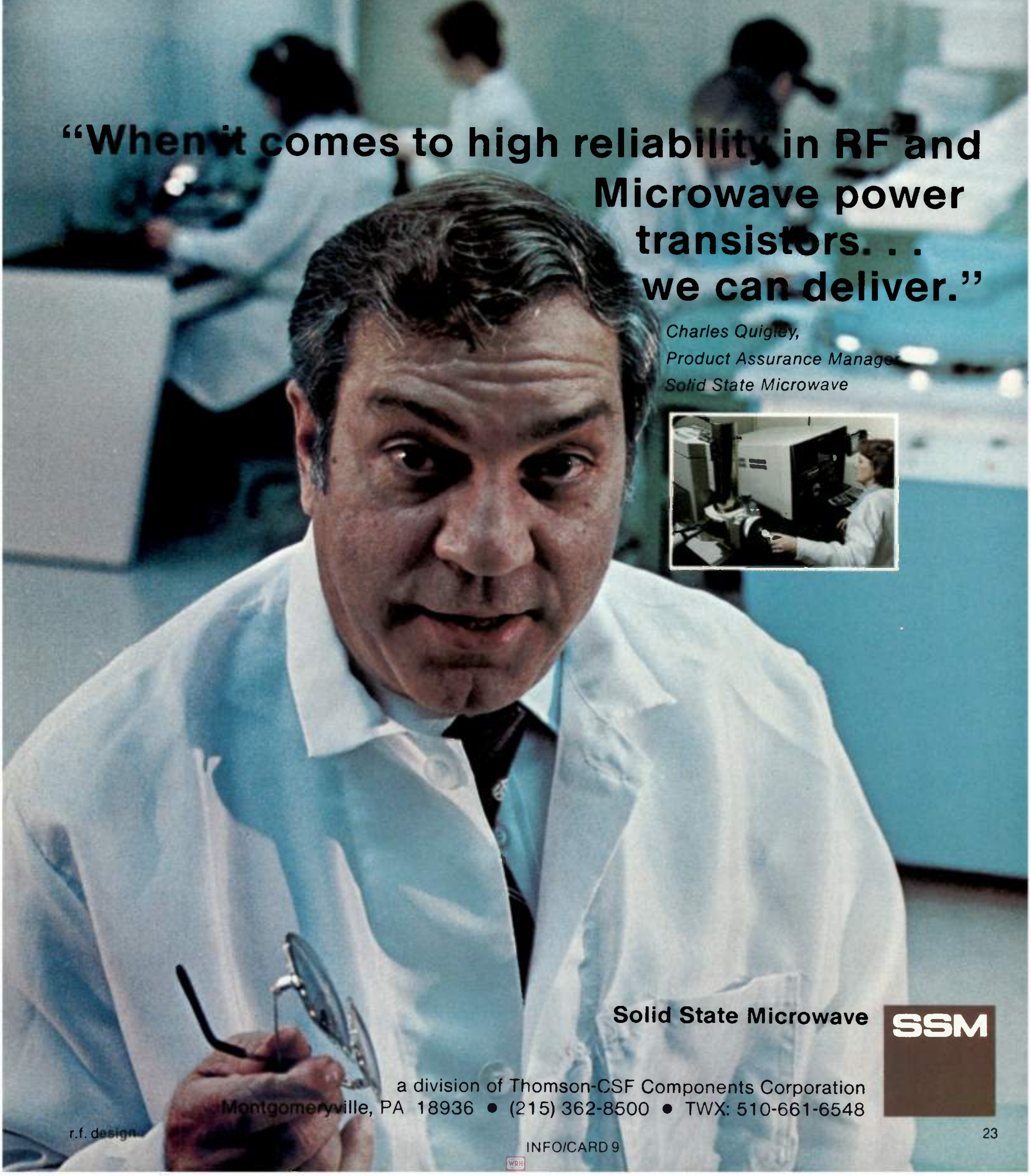

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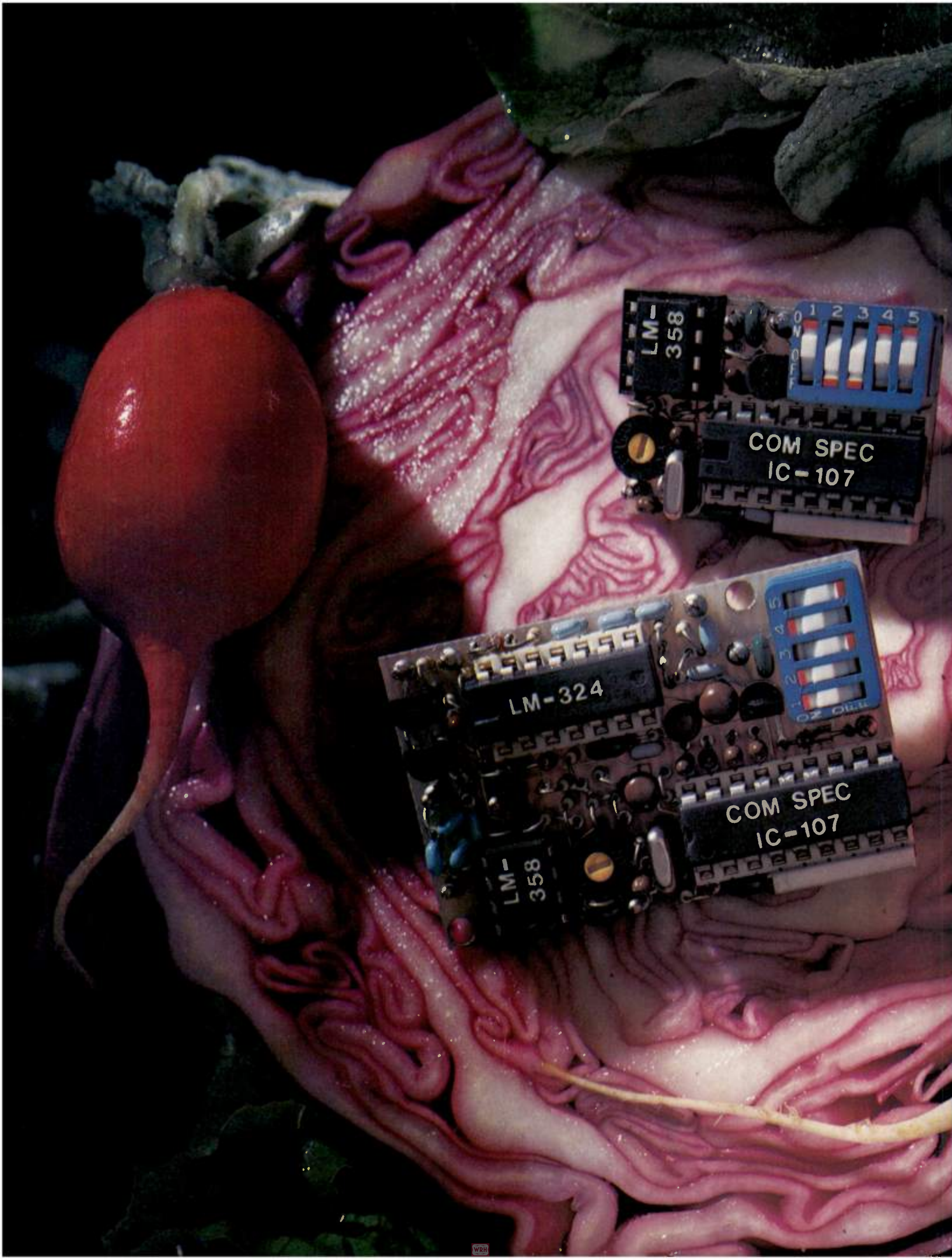
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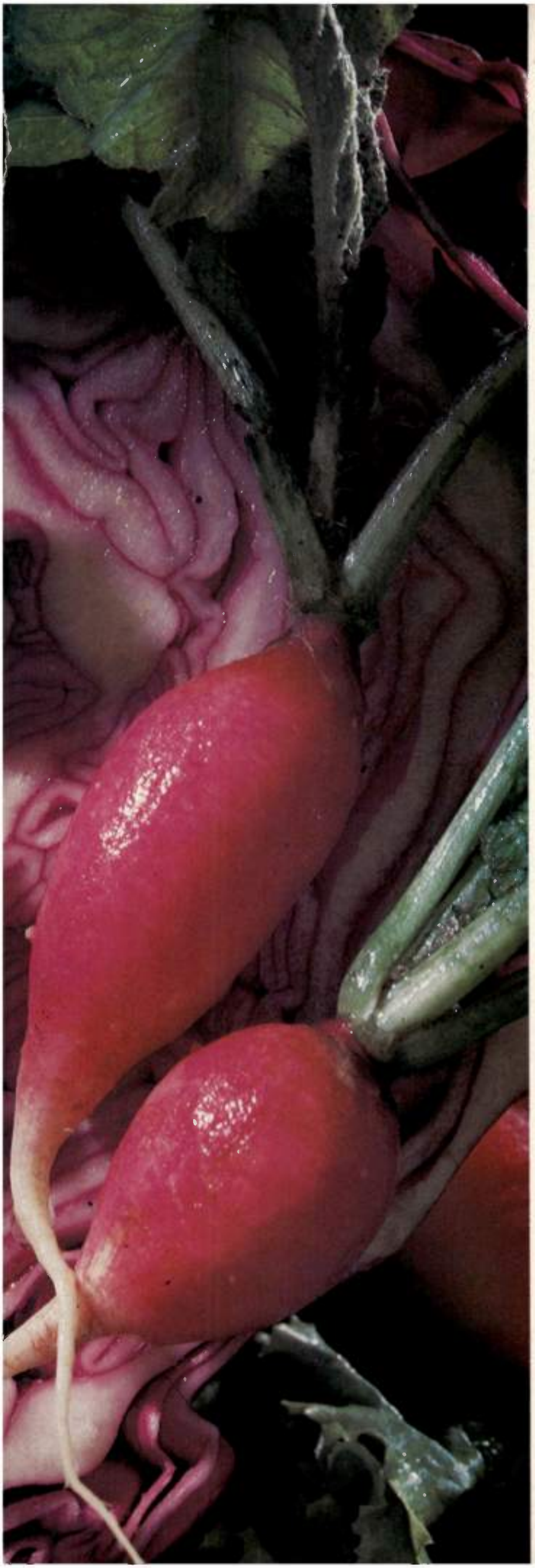

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74.4 WA	97.4 ZB	127.3 3A	167.9 6Z
77.0 XB	100.0 1Z	131.8 3B	173.8 6A
79.7 SP	103.5 1A	136.5 4Z	179.9 6B
82.5 YZ	107.2 1B	141.3 4A	186.2 7Z
85.4 YA	110.9 2Z	146.2 4B	192.8 7A
88.5 YB	114.8 2A	151.4 5Z	203.5 M1

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1500	852	1477	1700	1950	2250	2500
2175	941	1633	1750	2000	2300	2550
2805			1800	2100	2350	

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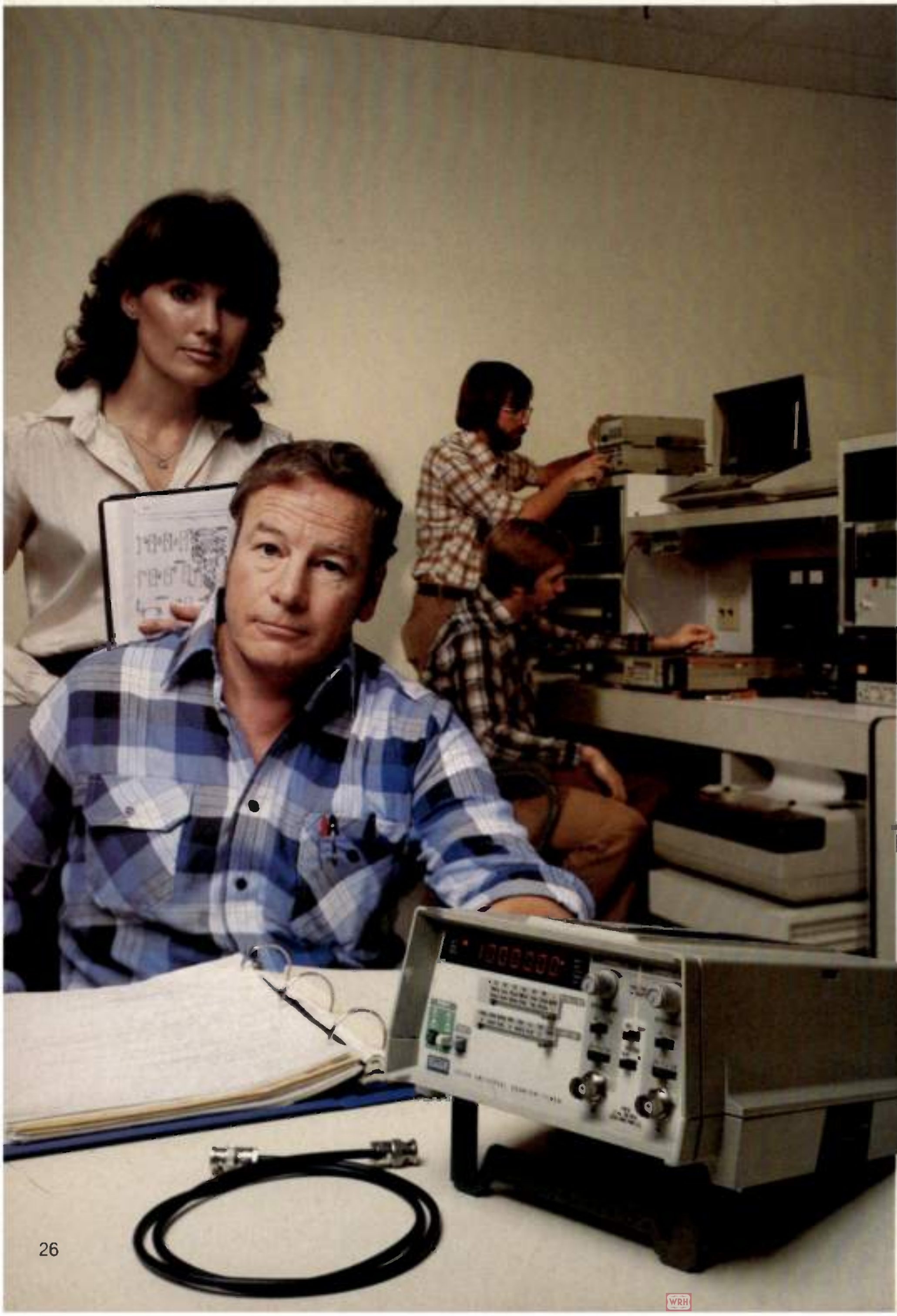

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Design of a Miniature Quadrature Coupler Using Lumped Capacitors And Transmission Lines

10% bandwidth miniature coupler analysis and Fortran 4 program design are presented.

Chen Y. Ho
Motorola, Inc.
Scottsdale, Az

Analysis and design of a miniature quadrature coupler consisting of lumped capacitors and transmission lines are presented. The coupler has a useful bandwidth of about 10 percent and is best suited for realization below 2 GHz where precise lumped capacitors are commercially available. The coupler is small in size since it requires only half the length of the conventional directional couplers. A computer program has been written for the analysis and design of this coupler and a list of the source program is included in the Appendix. Experimental results of a coupler design based on the design equations derived in this paper are very close to the theoretical computations.

Derivation of Design Equations

The quadrature coupler under investigation has the configuration shown in Figure 1, where two pieces of transmission lines with characteristic impedance Z_0 and electrical length θ_0 at a center frequency of f_0 are connected at both ends by a pair of lumped capacitors C. The impedance level of the coupler is also Z_0.

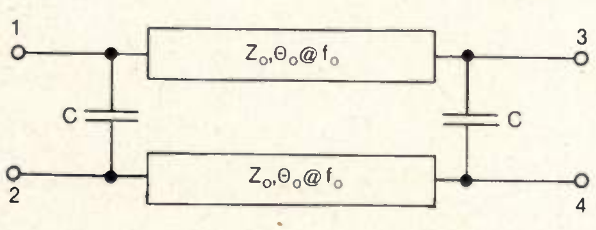


Figure 1. Coupler Configuration.

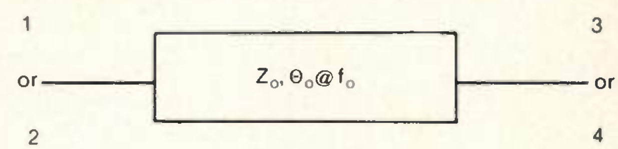


Figure 2a: Even-mode Equivalent Circuit.

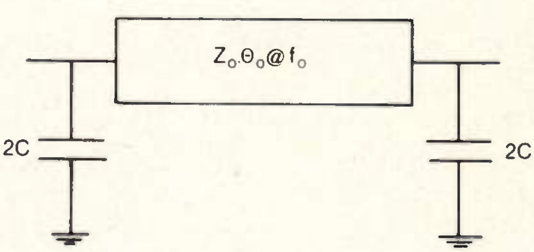


Figure 2b: Odd-mode Equivalent Circuit.

The analysis of such a coupler follows the even-mode and odd-mode approach[1] for symmetrical networks. For the even-mode excitation, the even-mode equivalent circuit is simply a piece of transmission line with characteristic impedance Z_0 as shown in Figure 2a. The even-mode ABCD matrix at a center frequency of f_0 is simply:

$$M+ = \begin{matrix} \cos(\theta_0) & jZ_0 \sin(\theta_0) \\ \hline j\sin(\theta_0)/Z_0 & \cos(\theta_0) \end{matrix} \qquad (1)$$

The even-mode reflection coefficient Γ_+ becomes:

$$\Gamma_+ = \frac{(AZ_o + B) - (CZ_o + D)Z_o}{(AZ_o + B) + (CZ_o + D)Z_o} = 0 \qquad (2)$$

and the even-mode transmission coefficient T_+ can be expressed as:

$$T_+ = \frac{2Z_o}{(AZ_o + B) + (CZ_o + D)Z_o}$$

$$= 1/(\cos(\theta_o) + j\sin(\theta_o)) \qquad (3)$$

Similarly, for odd-mode excitation, the odd-mode equivalent circuit is shown in Figure 2b. The odd-mode ABCD matrix becomes:

$$M_- = \begin{bmatrix} 1 & 0 \\ j2w_oC & 1 \end{bmatrix} X \begin{bmatrix} \cos(\theta_o) & j\sin(\theta_o)Z_o \\ j\sin(\theta_o)/Z_o & \cos(\theta_o) \end{bmatrix} X \begin{bmatrix} 1 & 0 \\ j2w_oC & 1 \end{bmatrix}$$

$$= \begin{bmatrix} \cos(\theta_o) - 2Cw_oZ_o\sin(\theta_o) & jZ_o\sin(\theta_o) \\ j[4Cw_o\cos(\theta_o) + \sin(\theta_o)/Z_o - 4C^2w_o^2Z_o\sin(\theta_o)] & \cos(\theta_o) - 2Cw_oZ_o\sin(\theta_o) \end{bmatrix} \qquad (4)$$

The odd-mode reflection coefficient Γ_- can be obtained from (4) as:

$$\Gamma_- = \frac{\left\{jZ_o\sin(\theta_o) - j(4Cw_o\cos(\theta_o) + \sin(\theta_o)/Z_o - 4C^2w_o^2Z_o\sin(\theta_o))Z_o^2\right\}}{\left\{2Z_o[\cos(\theta_o) - 2Cw_oZ_o\sin(\theta_o)] + jZ_o\sin(\theta_o) + j[4Cw_o\cos(\theta_o) + \sin(\theta_o)/Z_o - 4C^2w_o^2\sin(\theta_o)Z_o]Z_o^2\right\}} \qquad (5)$$

and the odd-mode transmission coefficient T_- becomes:

$$T_- = \frac{2Z_o}{\left\{2Z_o(\cos(\theta_o) - 2Cw_oZ_o\sin(\theta_o)) + jZ_o\sin(\theta_o) + j[4Cw_o\cos(\theta_o) + \sin(\theta_o)/Z_o - 4C^2w_o^2Z_o\sin(\theta_o)]Z_o^2\right\}} \qquad (6)$$

Since $\Gamma_+ = 0$, the total reflection at input port becomes 0 if $\Gamma_- = 0$, which implies that:

$$Cw_oZ_o^2\cos(\theta_o) = C^2w_o^2Z_o^3\sin(\theta_o)$$

$$\text{or} \quad \tan(\theta_o) = 1/(w_oCZ_o) = 1/(2\pi f_oCZ_o) \qquad (7)$$

Substituting (7) into (6), we have

$$T_- = \frac{2Z_o}{2Z_o(-Cw_oZ_o\sin(\theta_o)) + 2jZ_o\sin(\theta_o)}$$

$$= \frac{1}{-\cos(\theta_o) + j\sin(\theta_o)} \qquad (8)$$

If (7) holds, then the coupler is matched at f_o and the transmission from port 1 to port 2, $T_{21} = (\Gamma_+ - \Gamma_-)/2 = 0$, is also perfectly isolated at f_o. Furthermore, the transmission from port 3 and port 4, T_{31} and T_{41}, can be obtained as:

$$T_{31} = (T_+ + T_-)/2 = -j\sin(\theta_o) \qquad (9)$$

$$T_{41} = (T_+ - T_-)/2 = \cos(\theta_o) \qquad (10)$$

which indicated that the phase of T_{41} leads that of T_{31} by 90 degrees at f_o. If we select $\theta_o = 45°$, $T_{31} = T_{41}$, and the coupler becomes a 3 dB coupler. In this case the value of the lumped capacitors are:

$$C = \frac{1}{2\pi f_oZ_o} \qquad (11)$$

Equations (7), (9) and (10) are the design equations for the quadrature coupler of Figure 1. For a 3 dB design, $\theta_o = 45°$ and (11) can be used to obtain the value of the lumped capacitors. For example, at 1 GHz and $\theta_o = 45°$, a 3 dB coupler requires C = 1000/(2xπx50) = 3.16 pF. These capacitors are commercially available, for instance, from Di-Cap or precise MOS or MIS capacitors, up to S-band. The 3 dB coupler in this design has only half the electrical length of the conventional directional coupler and can be used effectively for applications requiring small areas.

Computed Frequency Responses

The frequency responses of the quadrature coupler of Figure 1 can be computed by using equations (2), (5), (3) and (6). For a 3 dB coupler design, the computed normalized frequency responses are shown in Figure 3a through 3c. As can be seen from these figures, the useful bandwidth of the coupler is about 10 percent. The source program for computing the frequency responses is listed in the Appendix.

Acknowledgment

The author wishes to express his gratitude to Jim Sturm of Motorola for the fabrication of the circuits.

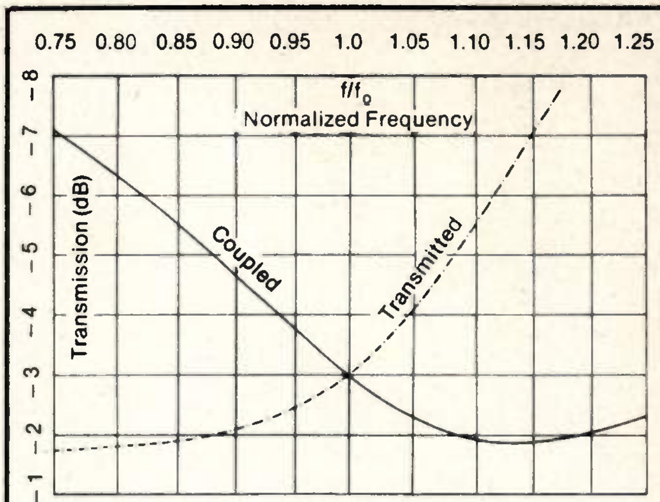


Figure 3a. Amplitude Response.

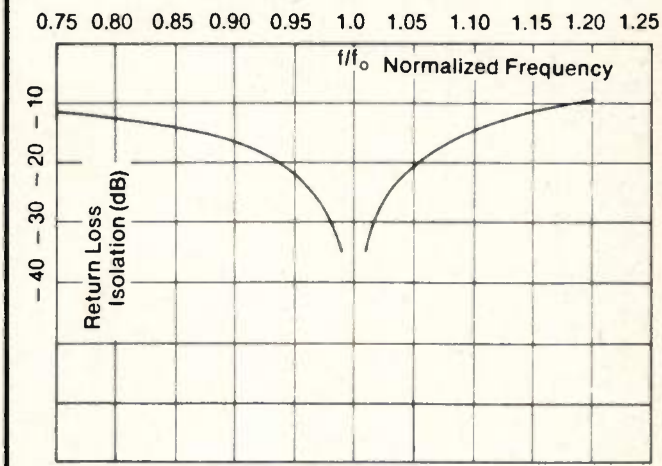


Figure 3b. Return Loss & Isolation.

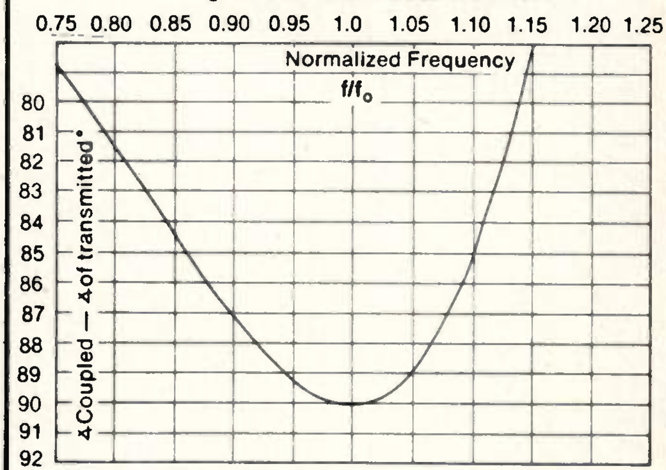


Figure 3c. Phase Response.

Experimental Results

A 3 dB coupler design at 0.667 GHz has been built and tested. The center frequency 0.667 GHz is selected because of the availability of 4.7 pF capacitors. The coupler is fabricated using standard MIC thick film techniques on a 25 mil thick alumina substrate. The layout of the coupler is shown in Figure 4 and the measured frequency responses are shown in Figure 5 which are in close agreement with the theoretical computations.

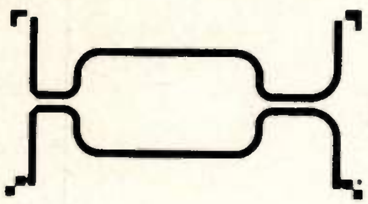


Figure 4. Experimental Coupler Layout.

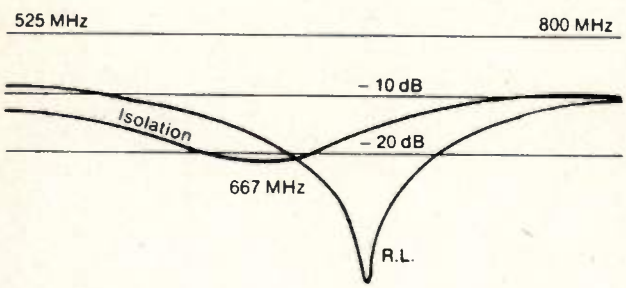


Figure 5a. Return Loss & Isolation Characteristics.

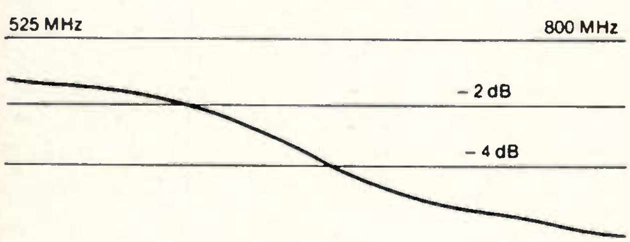


Figure 5b. Transmitted Port Characteristics.

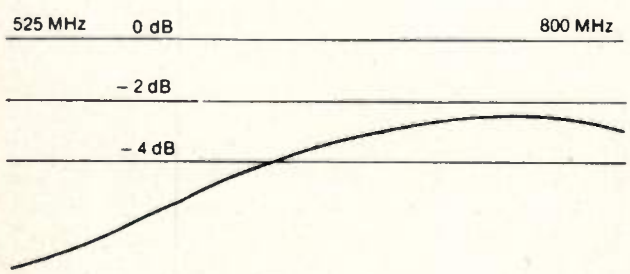


Figure 5c. Coupled Port Characteristics.

Appendix

```
C
C
C       THIS PROGRAM WILL DO THE DESIGN AND THE ANALYSIS OF END-COMPENSATED
C       QUADRATURE COUPLER WHICH CONSISTS OF COUPLED LINES OF LENGTH THETA
C       AND TWO CAPACITORS LOCATED AT BOTH ENDS OF THE COUPLER
C
C       FOR INPUT FORMAT:
C                       ZOE = EVEN MORE IMPEDANCE
C                       ZOO = ODD MODE IMPEDANCE
C                       THETA = ELECTRICAL LENGTH OF THE COUPLER.
C                       ZO  = CHARACTERISTIC IMPEDANCE
C                       CAP = CAPACITANCE IN THE COMPENSATION (PF)
C                       F1  = LOWER FREQUENCY IN GHZ
C                       F2  = UPPER FREQUENCY IN GHZ
C
C
        COMPLEX A,B C,D,YC,GAMMA(2),TRANS(2)
        COMPLEX CJ
        PI = 3.14159265359
        CJ = CMPLX(0.,1.)
   28   READ (105,11) ZOE, ZOO, THETA    ,ZO,CAP,F1,F2
   11   FORMAT(8F10.4)
        IF (ZOE-0.5) 21,21,22
   22   WRITE (108,12) ZOE, ZOO, THETA    ,ZO,       F1,F2
   12   FORMAT(1H1,4X,8HEVEN Z = ,F10.4/,5X,8H DDD Z = F10.4/,
       + 5X,8HTHETA    = ,F10.4/,
       + 5X,8HZO       = ,F10.4/,
       + 5X,8HF LOW    = ,F10 4, 4H GHZ/,5X 8HF HIGH = ,F10.4,4H GHZ)
        FO = 0.5 * (F1 + F2)
        DF = (F2 – F1)/50.
        IF (CAP – 1.E – 2) 16,16,17
   16   CAP = 1000./(2. * PI * FO * ZOO * TAN(THETA * PI/180.))
   17   WRITE (108,14) CAP
   14   FORMAT (5X,8HCAPAC. = ,F10.4,3H PF//)
        ZOE = ZOE/ZO
        ZOO = ZOO/ZO
        WRITE (108,13)
   13   FORMAT(//,7X,4HFREQ,8X,4HR.L.,8X,4HCOUP,8X,5HTRANS,5X,9HISOLATION,
       + 5X,5HPHASE,4X,11HDIRECTIVITY,/)
        DO 23  I = 1,51
        TI = I
        F = F1 + (TI – 1.) * DF
        TT = (F/FO) * (PI/2.)
        T = TT * THETA/90.
        CX = COS(T)
        SX = SIN(T)
        DO 24 J = 1,2
        IF (J – 1) 25,25,26
   25   A = CX
        B = CJ * SX * ZOE
        C = CJ * SX/ZOE
        D = CX
        GO TO 27
   26   YC = CJ * 2. * PI * F * CAP/1000.  * ZO * 2.
        A = CX + CJ * SX * ZOO * YC
        B = CJ * SX * ZOO
        C = 2. * CX * YC + CJ * SX/ZOO  +  CJ * YC * YC * SX * ZOO
        D = CX + CJ * SX * ZOO * YC
   27   GAMMA(J) = (A + B – C – D)/(A + B + C + D)
        TRANS(J) = 2./(A + B + C + D)
   24   CONTINUE
        TEST = CABS(0.5 * (GAMMA(1) + GAMMA(2)))
        IF (TEST – 1.E – 5) 31,31,32
   31   RL =  – 100.
        GO TO 33
   32   RL = 20  * ALOG10(TEST)
   33   COUP = 20. * ALOG10(CABS(0.5 * (TRANS(1) – TRANS(2))))
        TEST = CABS(0.5 * (GAMMA(1) – GAMMA(2)))
        IF (TEST – 1.E – 5) 34,34,35
   34   XISO =  – 100.
        GO TO 36
   35   XISO = 20. * ALOG10(TEST)
   36   DIRT = 20. * ALOG10(CABS(0.5 * (TRANS(1) + TRANS(2))))
        RE = REAL(0.5 * (TRANS(1) – TRANS(2)))
        RI = AIMAG(0.5 * (TRANS(1) – TRANS(2)))
        PH1 = ATAN2(RI,RE) * 180./PI
        RE = REAL(0.5 * (TRANS(1) + TRANS(2)))
        RI = AIMAG(0.5 * (TRANS(1) + TRANS(2)))
        PH2 = ATAN2(RI,RE) * 180./PI
        PHASE = PH1 – PH2
        DD = COUP – XISO
        WRITE (108,15)F,RL,COUP,DIRT,XISO,PHASE,DD
   15   FORMAT (8F12.4)
   23   CONTINUE
        GO TO 28
   21   STOP
        END
```

Conclusion

This paper has presented the design equations for the quadrature coupler of Figure 1. The coupler is small in size although it's useful bandwidth is limited to 10 percent. The experimental results of a 3 dB coupler are very close to the theoretical computations. A computer program for the analysis of the coupler is also included in the Appendix.

Reference

1. Reed, J and Wheeler, G.J, "A Method of Analysis of Symmetrical Four-Port Networks," *IRE Trans. On MTT*, vol. 4, pp. 246-252, Oct. 1956.

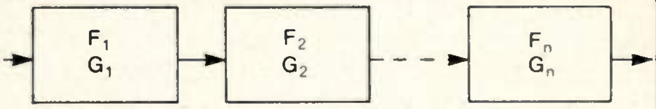
Noise Figure Calculation

The relative importance of the noise contribution of "downstream" components in cascaded networks is detailed.

Andrzej B. Przedpelski
A.R.F. Products, Inc.
Boulder, Colorado

We are all familiar with the equation for the overall noise figure of cascaded networks:[1]

The calculation of the overall noise figure is quite tedious and, therefore, is usually restricted to the first few stages of a typical RF receiver (RF amplifiers, mixer, and first IF stage). This can be quite misleading, as shown in the following example.



F = Stage NF
G = Stage gain

$$F = F1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots + \frac{F_n - 1}{G_1 G_2 \cdots G_{n-1}}$$

Figure 1. Cascaded Networks Noise Figure.

Figure 2 shows the basic block diagram of an FM satellite receiver designed for very large dynamic range. Using the usual approach (considering only the RF amplifier, filter, mixer, and first IF stage) an overall noise figure of 7.3 dB is obtained. Actually, however, the noise figure is 7.8 dB. All stages, in this case, have to be considered, even the first audio stage. This involves a considerable amount of calculation, which can be greatly simplified by using the program shown in Table I. This program is written for the HP-19/29 (use R/S in place of PRx for the HP-29), but

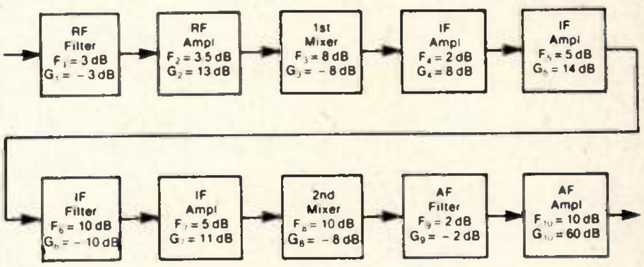


Figure 2. Special Purpose Satellite Receiver.

Noise Figure Calculation
Table I (Using HP 19/29)

Step	Instructions	Input Data/Units	Keys		Output Data/Units
1	Enter Program				
2	Store Values of F_x and G_x (Both in dB). Last Entry Should Be a G_{n-1} In An Odd Register	F_1	STO	3	
		F_2	STO	4	
		G_1	STO	5	
		F_3	STO	6	
		G_2	STO	7	
		• • • •			
		G_{n-1}	STO	2n + 1	
3	Store 0 in Next Register	0	STO	2n + 2	
4	Store 3 In Register 0	3	STO	0	
5	Calculate Noise Figure		GSB	0	F* F (dB)
6	If Desired, Change Any Value of F_x or G_x				
7	Store 3 In Register 0	3	STO	0	
8	Calculate New Noise Figure		GSB	0	F* F (dB)
	*First The Individual Stage Contributions Will Be Shown, Then The Overall Noise Figure				

Registers

0	3	1		2		3	F_1	4	F_2	5	G_1	6	F_3	7	G_2	8	F_4	9	G_3
10	F_5	11	G_4	12	F_6	13	G_5	14	F_7	15	G_6	16	F_8	17	G_7	18	F_9	19	G_8
20	F_{10}	21	G_9	22	0														

Of Cascaded Networks

can be modified for other similar calculators.

First, the noise figure of the first stage (F1) is converted from dB to a ratio (steps 2 through 8), then the next noise figure (F2) and the first gain (G1) are converted to a ratio, and the second term of the equation of Figure 1 is calculated (steps 11 through 32). When the calculation of the second term of the equation is finished, the process is repeated for all the remaining terms, storing the results in register 2 (step 33), until all the stored values of F and G are used up and 0 is reached in the next register (R20 in this example). The incrementing and calculating stops, and the result (the summation in storage register 2) is given (step 37). This total noise figure is then converted to dB (steps 38 to 41) and the final result, in dB, is obtained in step 42.

The readouts, in our example, are:

RF Filter	1.9953
RF Ampl	2.4716
Mixer	0.5310
IF Ampl	0.3690
IF Ampl	0.2162
IF Filter	0.0358
IF Ampl	0.0861
Mixer	0.0285
AF Filter	0.0117
AF Ampl	0.2846
Total	6.0297
Total, in dB	7.8030

It is evident that all the stages have some appreciable contribution to the final noise figure. Changing, for instance, the noise figure of the audio amplifier (usually not considered in noise figure calculations) from 10 dB to 20 dB, changes the overall receiver noise to 9.5 dB, with the following readouts:

	1.9953
	2.4716
	0.5310
	0.3690
	0.2162
	0.0358
	0.0861
	0.0285
	0.0117
	3.1307
Total	8.8758
Total, in dB	9.4821

The effect of changing any of the gains or noise figures on the overall noise figure can be easily evaluated by changing the desired storage register.

To obtain the correct answer the following should be remembered:

- enter gains and noise figures into the storage registers in dB, since this is how these values are usually given,
- start with F_1 in storage register R3 and enter the values consecutively, as shown, until the last gain (G_{n-1}) is entered and then enter "0" in the next storage register,

Step	Key Entry	Key Code
001	(g) LBL 0	25 14 00
	RCLi	55 12
	1	01
	0	00
	÷	61
	(g) 10^x	25 33
	STO 1	45 01
	PR_x	65
	1	01
010	STO 2	45 02
	(g) LBL 1	25 14 01
	ISZ	25 55
	RCLi	55 12
	(g) x = 0	25 61
	GTO 2	14 02
	1	01
	0	00
	÷	61
	(g) 10^x	25 33
020	1	01
	−	31
	ISZ	25 55
	RCLi	55 12
	1	01
	0	00
	÷	61
	(g) 10^x	25 33
	RCL 2	55 02
	x	51
030	STO 2	45 02
	÷	61
	PRx	65
	STO + 1	45 41 01
	GTO 1	14 01
	(g) LBL 2	25 14 02
	RCL 1	55 01
	PRx	65
	(f) log	16 33
	1	01
040	0	00
	x	51
	PRx	65
	(g) RTN	25 13

- always store "3" in register R0 before every calculation.

While the above example may be extreme, it is always a good practice to consider all possible sources of contribution to the overall noise figure to reduce the possibility of unpleasant surprises. □

Reference

1. Ghausi, Mohammed Shuaib: Principles and Design of Linear Active Circuits, McGraw-Hill Book Company, New York, N.Y., 1965.

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3	23	43	63	83	103	123
4	24	44	64	84	104	124
5	25	45	65	85	105	125
6	26	46	66	86	106	126
7	27	47	67	87	107	127
8	28	48	68	88	108	128
9	29	49	69	89	109	129
10	30	50	70	90	110	130
11	31	51	71	91	111	131
12	32	52	72	92	112	132
13	33	53	73	93	113	133
14	34	54	74	94	114	134
15	35	55	75	95	115	135
16	36	56	76	96	116	136
17	37	57	77	97	117	137
18	38	58	78	98	118	138
19	39	59	79	99	119	139
20	40	60	80	100	120	140

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1	21	41	61	81	101	121
2	22	42	62	82	102	122
3	23	43	63	83	103	123
4	24	44	64	84	104	124
5	25	45	65	85	105	125
6	26	46	66	86	106	126
7	27	47	67	87	107	127
8	28	48	68	88	108	128
9	29	49	69	89	109	129
10	30	50	70	90	110	130
11	31	51	71	91	111	131
12	32	52	72	92	112	132
13	33	53	73	93	113	133
14	34	54	74	94	114	134
15	35	55	75	95	115	135
16	36	56	76	96	116	136
17	37	57	77	97	117	137
18	38	58	78	98	118	138
19	39	59	79	99	119	139
20	40	60	80	100	120	140

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Electronic Test Accessory Catalog

ITT Pomona Electronics has published its 1980 catalog of electronic test accessories, 100 pages with three color photos, 562 black-and-white photos, and 115 drawings. Products covered include banana plugs, jacks and patch cords; phone tip jacks, plugs and connecting cords; test clips, probes, and holders; binding posts, black boxes and sockets; molded patch cords, cable assemblies, test socket adaptors; 3/4" spaced molded accessories, molded test leads and connecting leads. Special features of the catalog are a product index, two pages of new products, conversion tables of temperatures, metric converstion chart, BNC and triaxial cable procedures, a cross index of connector mil numbers, and a quantity price discount schedule.

Contact Carl W. Musarra ITT Pomona Electronics, 1500 East Ninth Street, Pomona, Calif. 91766. (714) 623-3463. INFO/CARD #136.

KJL Connector Catalog

ITT Cannon Electric has published a revised catalog on the KJL and KJ series of miniature circular connectors qualified to MIL-C-38999. Containing 32 photographs, 96 drawings, and 36 charts, catalog KJL/KJ-J5 comes with complete information on ordering, materials, finishes, electrical data and mechanical features, and contact arrangements. In the two series are wall mounting, straight plug, box mounting, and jam nut receptacle. A full page is devoted to test data — durability, insulation resistance, vibration, shock, thermal shock, altitude immersion, solvent immersion, corrosion, dynamic salt spray, and temperature durability.

Contact ITT Cannon Electric, 666 East Dyer Road, Santa Ana, Calif. 92702. (714) 557-4700. INFO/CARD #135.

CSC Solderless Stars In Sams Scope Sensation

Oscilloscope Applications and Experiments, the book, authored by Edward M. Noll and published by Howard W. Sams & Co., Inc. was written primarily for the novice user of oscilloscopes, also has a great deal of information that more advanced electronic engineers and technicians will find valuable. The informational content of the text is enhanced by the inclusion of a substantial number of practical experiments, most of which require simple peripheral circuits. Author Noll has built these circuits on CSC's popular Quick Test solderless breadboard elements, which he calls out by number in the parts list accompanying each experiment. These projects include counters, timers, dividers, gates, inverters, ramp generators, operational amplifiers, differential amplifiers, sine wave generators, square wave generators, two-tone generators, pulse position modulators, pulse generators, pulse width modulators and more. The book carries a suggested U.S. resale of $8.95; it is a 244-page volume.

Contact CSC, 351 California St., San Francisco, Calif. 94104. (415) 421-8872. INFO/CARD #134.

New Instrument Catalog

Nearly 100 advanced design, state-of-the-art, electronic test instruments are being featured in a new, 16-page, short form availabilities catalog recently published by The Instrument Mart of Great Neck, N.Y. The publication details key features and specifications of the test equipment produced by 14 separate manufacturers, thus providing a choice for engineers and technicians to meet most every conceivable, non-esoteric, electronic test and measuring requirement. Product groups offered in The Instrument Mart catalog include: oscilloscopes, digital multimeters, frequency counters, function and pulse generators, audio analyzers, NTSC generators, power supplies, digital thermometers and more. Manufacturers represented in the publication are: John Fluke Manufacturing, Philips Test & Measuring, Gould, Leader, B&K Dynascan, Beckman, Weston, Simpson and others. The Instrument Mart was established in 1970 and until recently, operated only in the Tri-State, New York Metropolitan region.

Contact The Instrument Mart, 295 Community Drive, Great Neck, N.Y. 11021. (516) 487-7430. INFO/CARD #133.

Instruments & Controls Shielding Techniques

Metex Corporation's Electronic Shielding Group has issued a new, six-page illustrated bulletin on instruments and controls shielding techniques provocatively titled "Up Your EMI Protection". Included are photos of such typical approaches as EMI gasketing material applied to an air filter; EMI shielded windows for meters, readouts, etc.; zip-on cable shielding; drawings identifying typical shielding and static discharge problem areas in equipment enclosure design; plus a chart of three Fourier spectra of EMI in a typical process control environment.

Contact Metex Corporation, Electronics Shielding Group, 970 New Durham Road, Edison, N.J. 08817. INFO/CARD #140.

1280-2560 MHz Frequency Doubler

Extending the high quality signal performance of synthesized signal generators can be an attractive alternative to buying another generator. The frequency range is doubled using this new HP11721A frequency doulber on the output.

Used with the HP8662A signal generator, the HP11721A frequency doubler extends the output frequency from 1280 MHz to 2560 MHz. CW and FM modulated signals reproduce with little distortion and with fairly predictable conversion loss. At drive levels greater than +10 dBm, for example, conversion loss is less than 15 dB. The HP11721A doubler can also be used with many other generators.

A new application note, AN 283-2, describes the operating performance of the HP11721A doubler and how it affects modulation, spectral purity, conversion loss, and output dynamic range.

Software routines are included for use in programmable applications of the HP Model 8662A Signal Generator.

Application Note 283-2, 'External Frequency Doubling of the 8662A Synthesized Signal Generator' is available at no cost.

Contact Inquiries Manager, Hewlett-Packard Company, 1507 Page Mill Road, Palo Alto, Calif. 94304. □

Plastic Made Electrically Conductive

R.P. Stromberg of Carbon Products Division of Union Carbide Corporation, examines simulated air scoop on the hood of an automobile mock-up, demonstrating treatment with carbon fiber to overcome radio frequency interference. The carbon fiber shielding is an integral part of the sheet molding compound from which the hood is formed. Contact Union Carbide Corporation, Carbon Products Division, 270 Park Avenue, New York, N.Y. 10017. Circle INFO/CARD #131.

400-512 MHz Cavity Filter

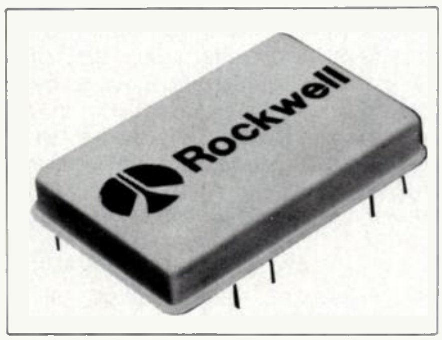
The model ACF400-512VD is an adjustable cavity filter designed for receive applications in the 400-512 MHz range. Typical insertion loss and 1 dB bandwidth are 0.2 dB and 8 MHz. Return loss is 30 dB (1.07 VSWR). Capacitive coupling is used to provide excellent low-frequency rejection with typical attenuation values as follows: at 300 MHz,

40 dB; at 200 MHz, 50 dB; at 100 MHz, 75 dB. The unit is constructed from heavy-wall copper tubing with all surfaces silver plated to ensure deterioration free performance. External surfaces are finished in grey enamel. Dimensions, excluding tuning screw, are 3-1/8" W X 3-3/16" H X 5" D. BNC connectors are standard. Other connectors or connector combinations are available.

Contact: Advanced Receiver Research, Box 1242, Burlington, Ct. 06013. INFO/CARD #128.

SAW Bandpass Filters

A standard product line of surface acoustic wave (SAW) bandpass filters for communications systems and radar IF systems has been announced by Rockwell International, Filter Products, Newport Beach, California. The SAW filters, with center frequencies of 70 MHz and 150 MHz, offer all the solid-state advantages of small size, low weight and high re-



liability. Rockwell SAW filters offer a flat group delay, controlled amplitude and time response, and phase linearity. Applications include radio frequency receivers, frequency synthesizers, channellized receivers and equalizers for digital data systems. Size of the filters is a maximum of 2.35" by 1.15" by .4".

Contact Rockwell International, Filter Products, 4311 Jamboree Road, Newport Beach, Calif. 92660. (714) 833-4324/4544. INFO/CARD #129.

Ovenized Frequency Standard

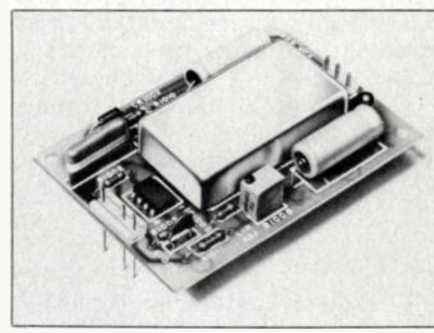
Greenray Industries now offers a low cost, highly stable, 5 MHz, sine wave output, ovenized frequency standard, GRI model YH-1050-1. The oscillator operates on a single +12 VDC supply. The unit maintains $\pm 5 \times 10^{-8}$ stability over a 0 to +65°C temperature range. It also has a guaranteed aging rate of $\pm 5 \times 10^{-9}$ per day, but typically only ages 1 to 3 x 10^{-9} per day. The size of the unit is 4.0"L x 2.0"W x 2.0" high.

Contact Donald R. Widener, 840 West Church Road, Mechanicsburg, Pa. 17055. (717) 766-0223. INFO/CARD #130.

Voltage Controlled Amplifier

A new low distortion, low noise, high bandwidth voltage controlled amplifier has been introduced by the Professional Products Division of dbx, Inc. The all-discrete design has a minimum gain-bandwidth product of 50 MHz, resulting in full audio bandwidth at 60 dB of gain with complete freedom from slew-rate related problems.

The gain versus control voltage characteristics of the dbx Model 2001 are an exponential function (20 dB/volt) allowing the user to easily and accurately program the amplifier gain directly in dB.



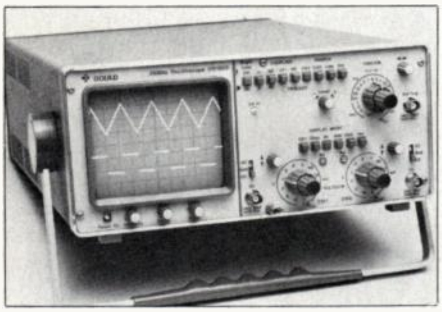
The dbx model 2001 is a "current in, current out" device. By choosing appropriate input and output resistors, the VCA can be optimized to fit the needs of the user. Like all dbx VCA's, it is a complete unit requiring a minimum amount of external components.

Contact the Professional Products Division, dbx, Inc., 71 Chapel Street, Newton, Mass. 02195. INFO/CARD #127.

Dual Trace 25 MHz Portable Oscilloscope

A new lightweight 25 MHz dual-trace oscilloscope, the OS1200, designed for analog and digital uses in both laboratory and field service applications has been developed by Gould Inc., Instruments Division. Weighing less than 17 lb., the unit is housed in a rugged metal enclosure 5-1/4 by 12 x 16-1/4 in. deep.

The Gould OS1200 features a full 5-in. rectangular CRT with internal graticule for optimum measurement accuracy. The unit's 14 ns risetime, 6 kV accelerating potential for increased definition and brightness, and the incorporation of signal delay make the instrument particularly useful for digital work involving narrow pulses with fast risetimes and low repetition frequencies. A 1-volt DC coupled Z-modulation input facilitates use with logic analyzer outputs.

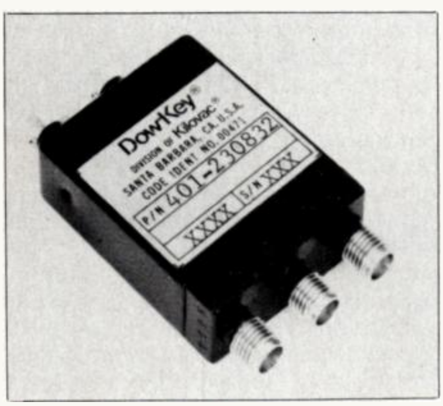


Basic timebase speeds range from 200 ns/cm to 1 s/cm with vernier control over a range of 2.5:1 (X10 expansion yields a maximum sweep rate of 20 ns/cm). The dual input channels have a maximum sensitivity of 2 mV/cm over the full 25 MHz bandwidth. Display modes include CH1 or CH2 alone, CH1 and CH2 alternate or chopped, CH1 ± CH2, and X-Y.

Contact Marketing Services, Gould Inc., Instruments Division, 3631 Perkins Avenue, Cleveland, Ohio 44114. (216) 361-3315. INFO/CARD #122.

High Frequency Relay

Dow-Key Division of Kilovac Corporation announces the availability of the model 401-230832 single-pole, double-throw coaxial relay. The new relay is designed for excellent RF performance in microwave systems to 18 GHz and for operation in severe environments. The 401-230832 is a failsafe relay remotely



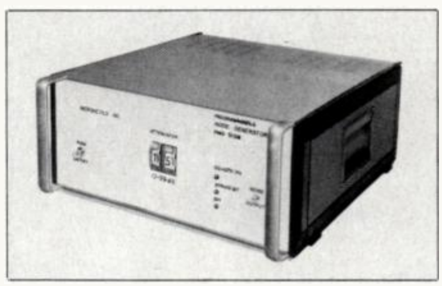
actuated by a 28 VDC, 475 ohm coil with a maximum operate time of 20 ms at 20°C. A balanced actuator mechanism provides very good tolerance to shock and vibration. Auxiliary contacts are provided for remote position indicator circuits.

The 401-230832 can carry 10 watts of RF power at 18 GHz with isolation greater than −60 dB, insertion loss less than −0.5 dB, and VSWR 1.5:1 maximum. At 1 GHz VSWR is less than 1.05:1, isolation is greater than −80 dB, insertion loss is less than −0.1 dB, and the maximum operating power is 75 W. At lower frequencies the 401-230832 is an excellent choice for high isolation with low VSWR and insertion loss. The maximum power capacity of the relay is 200 Watts up to 200 MHz. Typical applications include switching of antennas, components, receivers, and instruments in radar communications.

Contact Jack Dysart, Product Line Supervisor, P.O. Box 4422, Santa Barbara, Calif. 93103. (805) 684-4560. INFO/CARD #121.

Programmable Broadband Solid State Noise Generators

Replacing conventional sweep generators, Micronetics' model PNG 5100 through 5110 series of noise generators feature full noise source capability from 10 Hz to 1 GHz in a compact instrument or rack-mounted panel, only 7 inches high.



Output is greater than +10 dBm (10 mW) across the given frequency range.

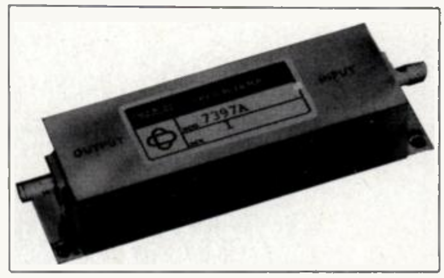
IEEE-488 interface bus remote programming includes on and off standby, external pulse and attenuation control, as well as local and remote capability.

Standard frequency bands are: 10 Hz-20 kHz, 10 Hz-100 kHz, 10 Hz-500 kHz, 100 Hz-3 MHz, 100 Hz-10 MHz, 100 Hz-25 MHz, 100 Hz-100 MHz, 1 MHz-200 MHz, 1 MHz-500 MHz, 10 MHz-1 GHz. Optional frequency bands are available up to 40 GHz.

Contact Micronetics Inc., 36 Oak Street, Norwood, N.J. 07648. (201) 767-1320. INFO/CARD #117.

Filter Aids Satellite Security

A high-frequency, quartz crystal band-pass filter, designed to enhance the performance of worldwide U.S. military satellite communications, has been developed by Damon Electronics, a division of Damon Corporation. Damon's Model 7397A filter, permits the expansion of military satellite communications. Capable of precision operation at frequencies as high as 150 magahertz, with a spurious-free bandwidth of 50 kilohertz nominal, the new Damon filter provides out-of-band spurious rejection of 60 dB



minimum, and ultimate rejection of 90 dB minimum, coupled with low insertion loss of 6 dB maximum. Gold-plated for protection against corrosive atmospheres and hermetically sealed, the device is designed for dependable, accurate performance in ambient temperatures ranging from −10°C to 80°C. Weighing 100 grams (3.5 ounces) and measuring 7.62 centimeters (three inches) in length, the Damon crystal filter is fitted with OSM connectors and is designed for use in single-conversion UHF radio receivers in both military and NASA earth-orbit satellites, as well as in ground communications stations.

Contact Damon Electronics, 80 Wilson Way, Westwood, Mass. 02090. (617) 449-0800. INFO/CARD #109.

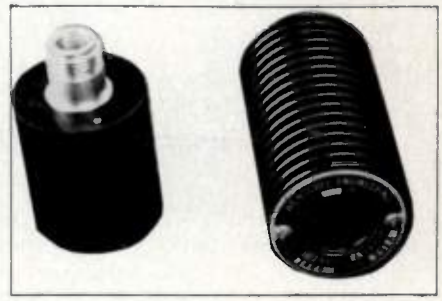
Hand-Held Transceiver And Test Equipment Antennas

RF Products announces a new line of telescoping and flexible whip antennas intended for use on VHF/UHF hand-hend transceivers or as RF pick up antennas on test equipment. Ideal for frequency counters and communication service monitors. Available with BNC, TNC, and PL-259 type connectors and with fixed right angle, adjustable angle, or straight (inline) whips. The Porta-Tenna is a multisection chrome-plated brass telescoping antenna. The Flex-Tenna utilizes an insulated flexible spiral stranded wire as the antenna whip. The Spring-Tenna is an insulated helical spring type antenna.

Contact RF Products, P.O. Box 33, Rockledge, Fla. 32955. (305) 631-0775. INFO/CARD #112.

50 Watt RF Terminations

The models 1426 and 1427 broadband coaxial terminations are for high-power applications from DC to 8 GHz and have a maximum average power input of 50 watts and 25 watts, respectively (5 kW peak for both models); excellent stability even when subjected to adverse environmental and overload conditions; and a low VSWR of 1.20 to 4 GHz and 1.30 to 8 GHz. Connectors



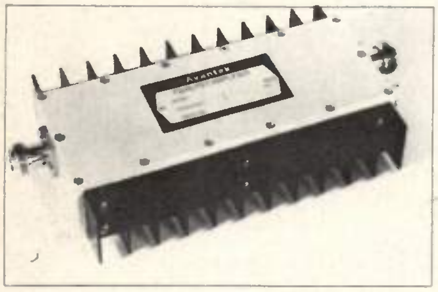
are Weinschel Engineering semi-precision type N male or female which mate with same type connectors per MIL-C-39012.

Contact Weinschel Engineering, Gaithersburg, Maryland 20760. (301) 948-3434. INFO/CARD #107.

50 to 2000 MHz GaAs FET Amplifiers

Avantek, Inc. is offering a series of GaAs FET amplifiers that cover the 50 MHz to 2000 MHz frequency range and feature up to +27 dBm output power (1 dB gain compression) with +37 dBm typical third-order intercept point. Other features include a choice of 18 or 28 dB standard gain (higher gain available on special order) with ±1.0 dB gain flatness, 6.0 dB noise figure and 2.0:1 maximum input and output VSWR in a 50 ohm system. These amplifiers, designated the APG-2050 Series, operate from unregulated power sources supplying +15 VDC nominal and 275 or 525 mA of current. APG-2050 Series amplifiers are useful as post-amplifiers to increase the output power capability of wideband sweep generators operating in the 50 to 2000 MHz frequency range. Their combination of wide dynamic range with low noise figure makes them a good choice as IF amplifiers in advanced receiving systems and instruments, particularly when they follow high LO-power mixers. These amplifiers may also be used as intermediate power amplifiers (IPAs) to drive the high power amplifiers in wideband jammers and other electronic defense applications.

These Avantek amplifiers are packaged in O-ring sealed aluminum cases with heat dissipating fins and are compact and lightweight. All amplification stages are actively biased GaAs FETs to insure wide dynamic range and good temperature stability. An integral DC-to-DC converter provides both positive and negative bias for



the GaAs FETs, which allows the source terminals to be positively grounded for electrical stability and effective heat dissipation. Negative feedback on each stage minimizes both input and output VSWR, improves gain flatness and eliminates the need for isolators.

APG-2050 Series amplifiers can be qualified to military specifications including MIL-E-16400, MIL-E-5400 and MIL-E-4158. They are capable of meeting the EMI conditions of MIL-STD-461 and are produced under a quality control/quality assurance program that meets MIL-Q-9858A requirements.

Contact Avantek, 3175 Bowers Ave., Santa Clara, Calif. 95051. (408) 249-0700. INFO/CARD #108.

Advertiser Index

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Forget everything you know about pulse generators.

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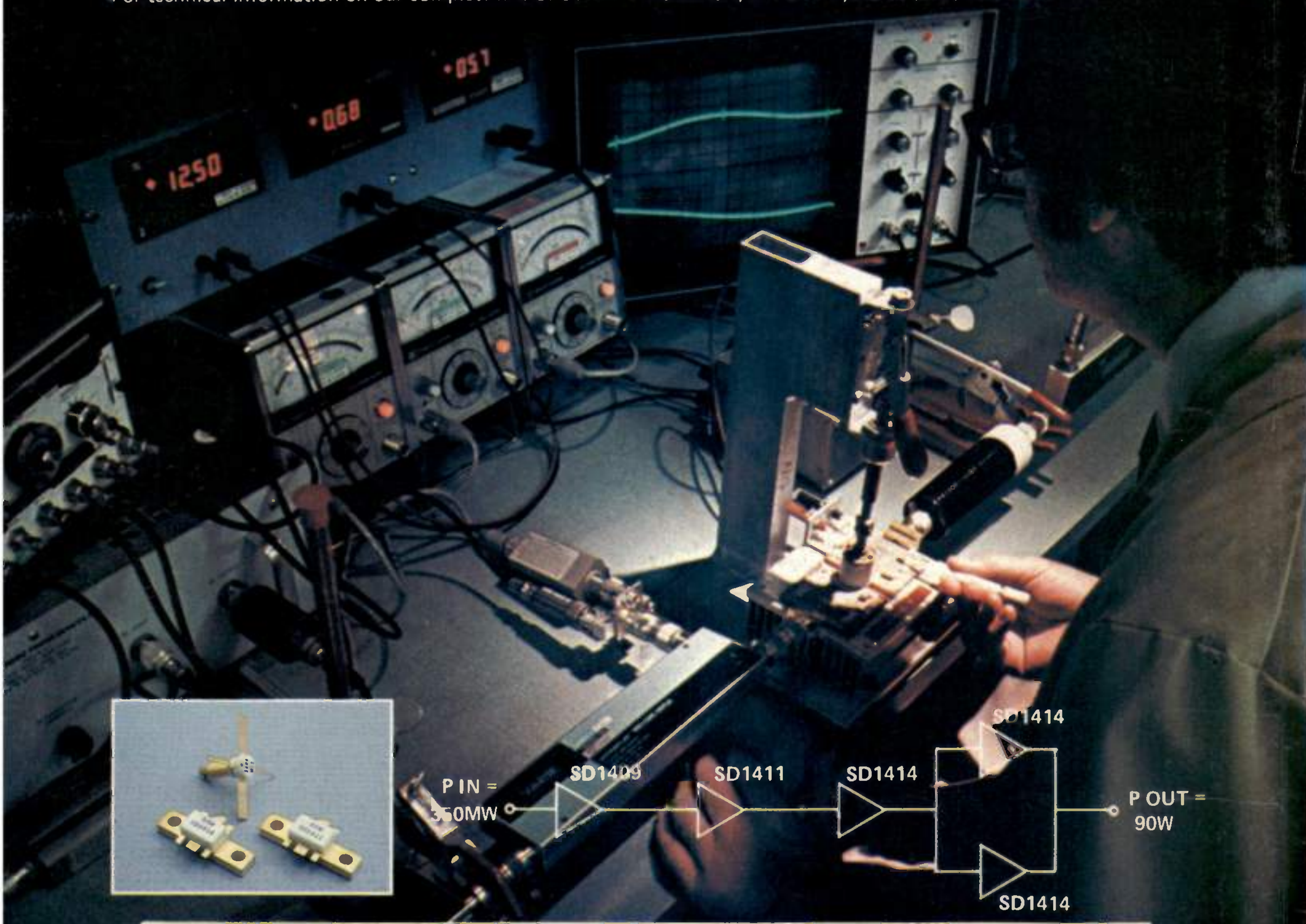
For the last two years Solid State Microwave has proven its leadership with *high volume production* of Microwave Power Transistors for the new 800MHz Land Mobile Industry. We have been involved in this new market from the beginning and have developed a sound, reliable, highly reproducible line of devices which have satisfied many large production contracts. When you are considering Microwave Power Transistors for your next 800MHz Land Mobile design, consider Solid State Microwave, the 800MHz *technology* and *production* leader.

For technical information on our complete line of 800MHz Transistors, call Bob Tyson at (215) 362-8500.



DEVICE TYPE	P OUT POWER OUTPUT WATTS	P IN INPUT POWER WATTS	PG POWER GAIN dB MIN	fo TEST FREQUENCY MHz	PACKAGE
SD1409	2.0	.35	7.6	870	XO72
SD1411	20	4.5	6.5	836	.230 6LFL
SD1414	50	15	5.2	836	.230 6LFL

V_{cc} (supply voltage: 12.5 volt)

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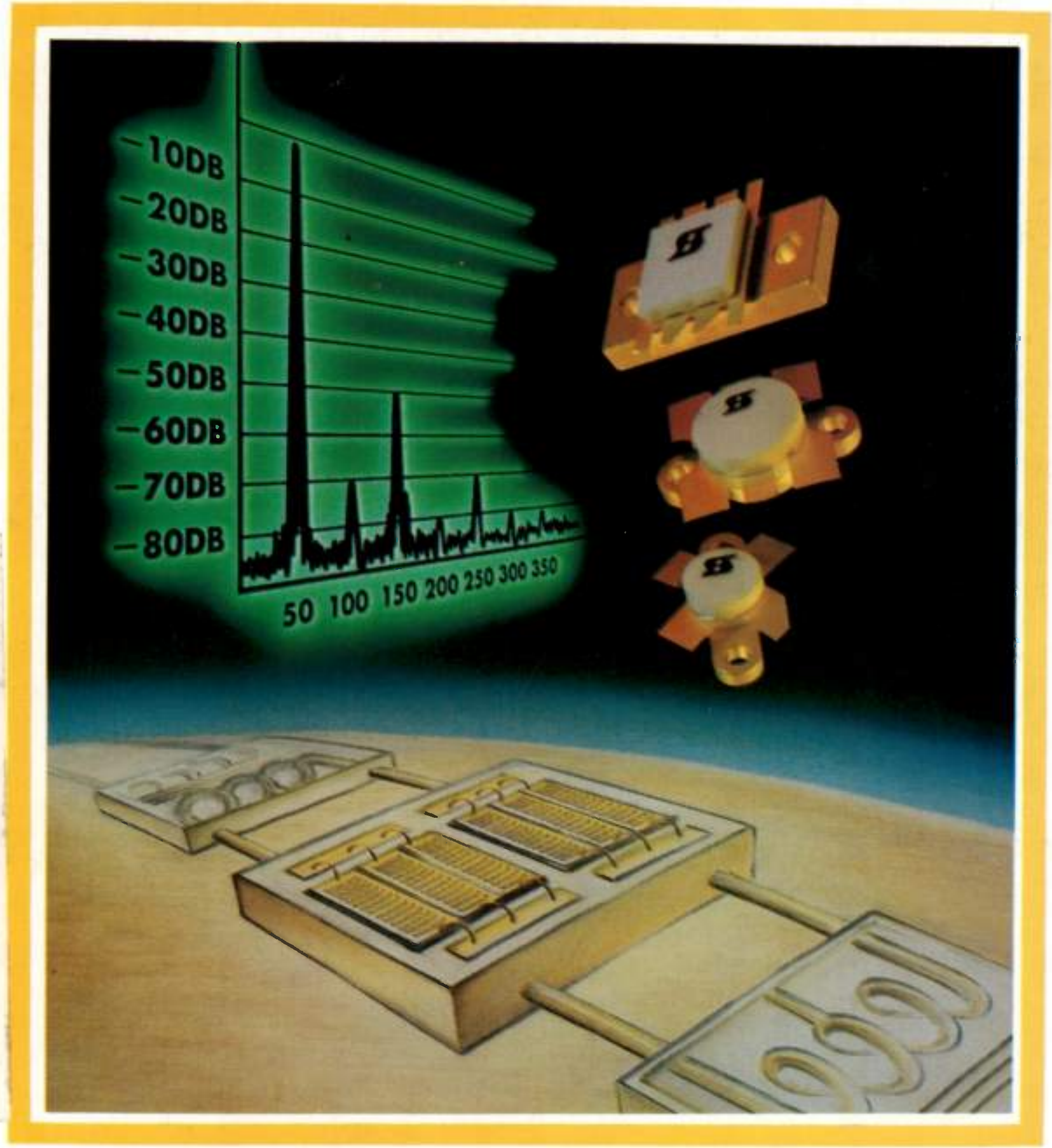
INFO/CARD 19

March 1980

A Cardiff Publication

rf design



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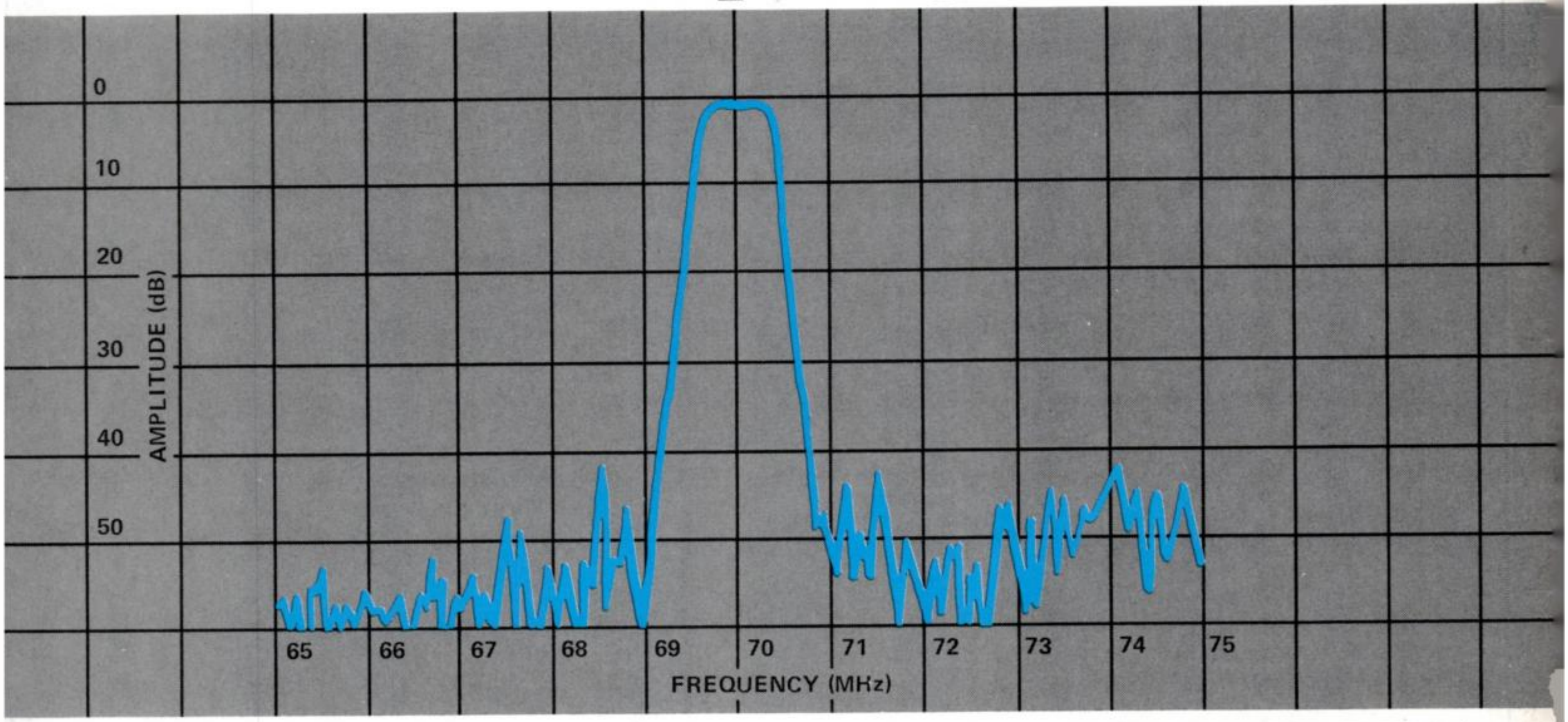

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r.f. design

March 1980

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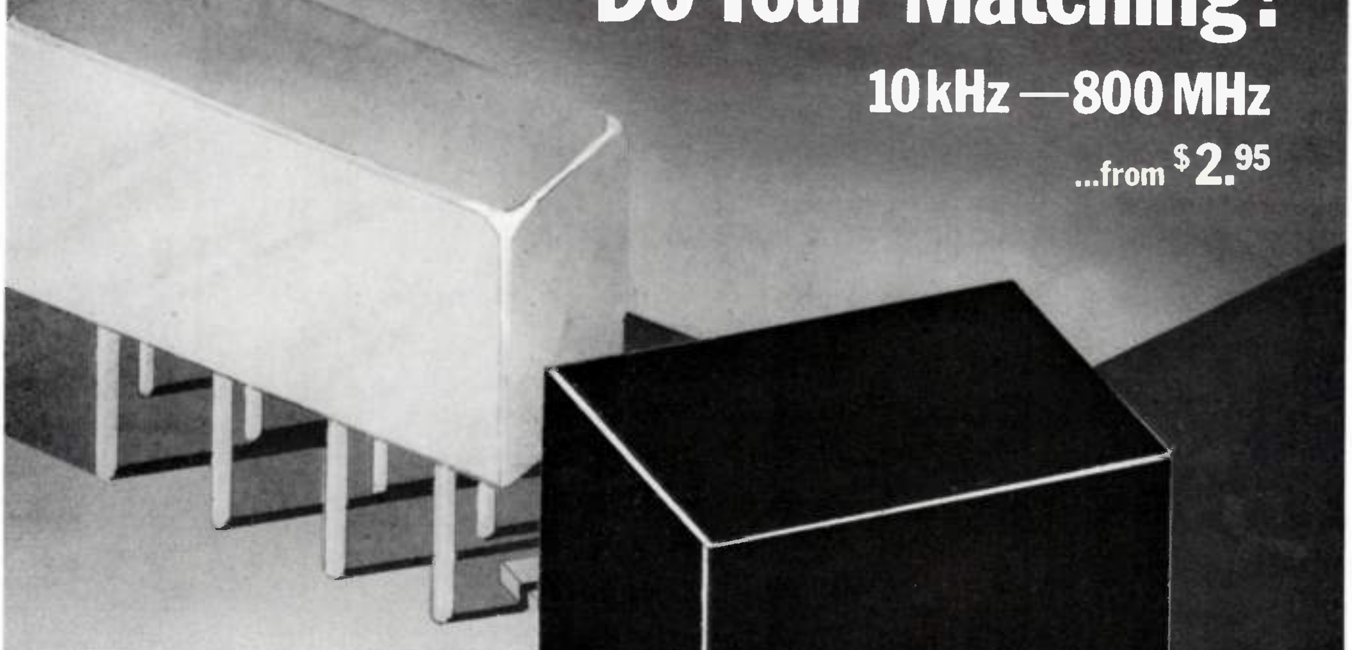

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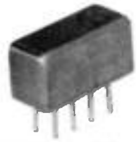

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TM04-1	4	.2-350	$4 95
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T4-2	4	.2-600	$3 45
TM04-2	4	.2-600	$5 95
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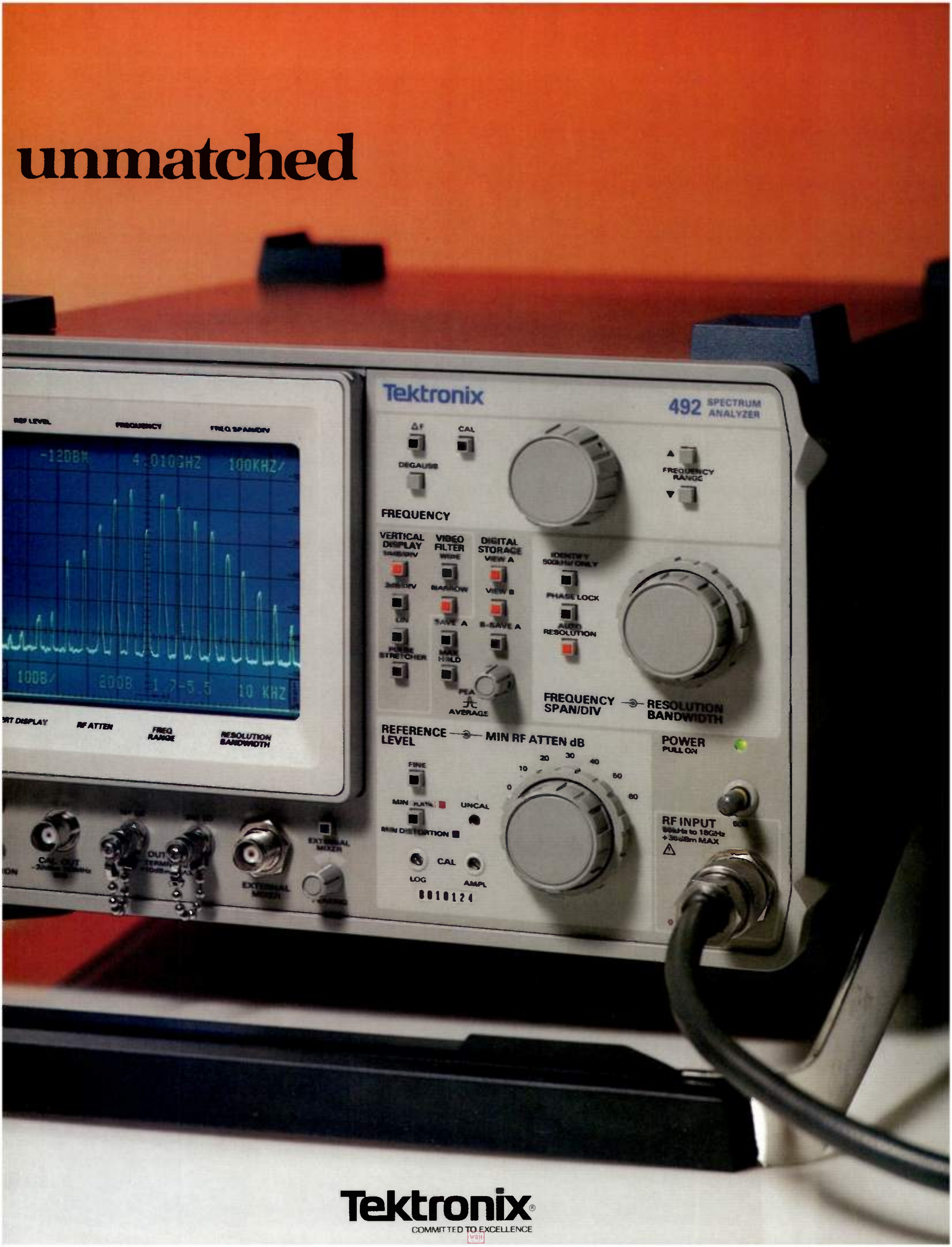
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We Have Come A Long Way (Or What is an RF Engineer?)

The RF Engineer is a hardy soul. He is a fine blend of intelligence, curiousity, creativity and most importantly stubborness. He has to be — for in what other engineering discipline would one spend hour after hour calibrating test equipment fixtures just to measure single point data on a device, stop, readjust a trimmer and start all over again?

Fifteen years ago that was a typical scene even in the large well-equipped engineering labs. Beg, borrow or steal a row of sweep generators, power meters, SWR indicators, detectors, couplers and a good oscilloscope, clear an area, catch one's breath (that equipment was heavy!) and start.

There we had it — fifteen hundred pounds of test equipment ready to pounce on this tiny little amplifier, filter or oscillator in which even the mating connectors were larger than the actual DUT. The aim was to measure its insertion loss, return loss and output power levels.

In 1966 Engineers rejoiced for their job became somewhat easier with the introduction of a vector voltmeter, an instrument combining several needed functions in one unit. (This is not to denigrate other fine pieces of equipment that existed at that time.)

It is now 1980 and a quick scan of a typical RF design lab reveals test equipment not even dreamt of by our hard working engineer of 15 years ago: oscilloscopes with 1000 MHz of video bandwidth, frequency counters with 11 digit precision and 10 MHz to 18 GHz continuous sweep scalar network analyzers (featured this month) all capable of "talking and listening" to each other as a result of an interface standard known as IEEE-488-1975.

We have come a long way since 1965, in many areas; in component technology, test equipment and foremost knowledge. The intriguing thought is what will the next 15, 30 or even 100 years hold? Inductively (logic i.e.) speaking it's mind boggling if we let ourselves dream.

In the words of the motto of my native state "excelsior" or Ever Upwards let us proceed . . .

(In retrospect that which has come before, i.e. the technical articles since volume one, number one of *r.f. design* is indexed in this issue.)

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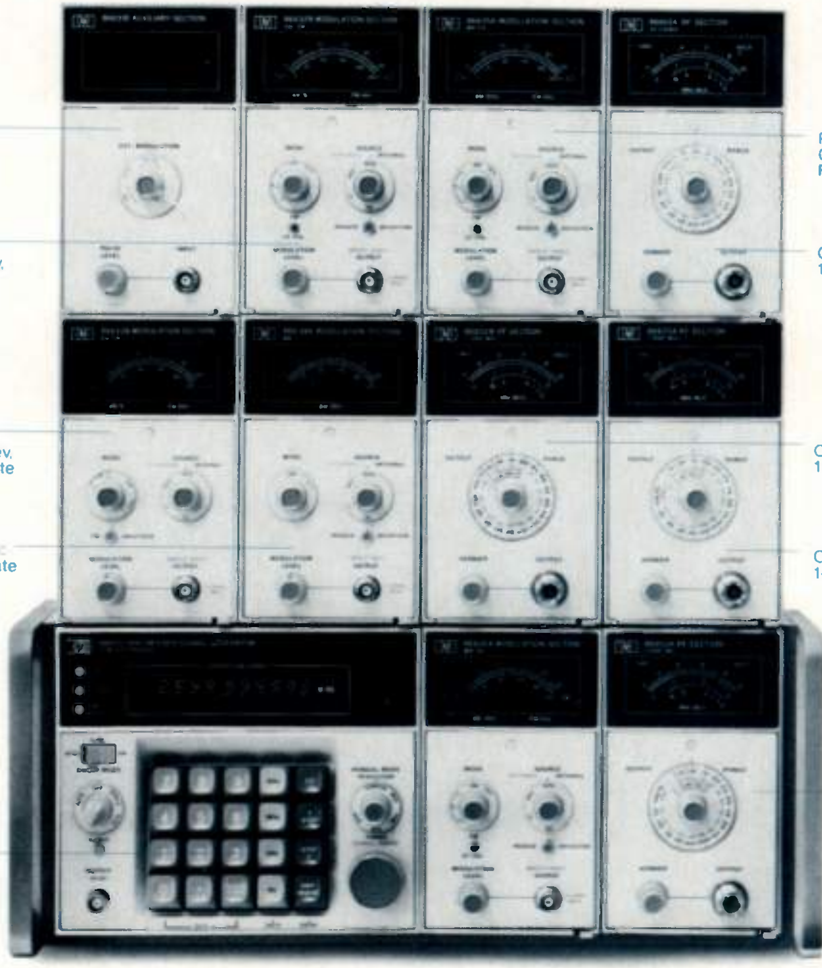

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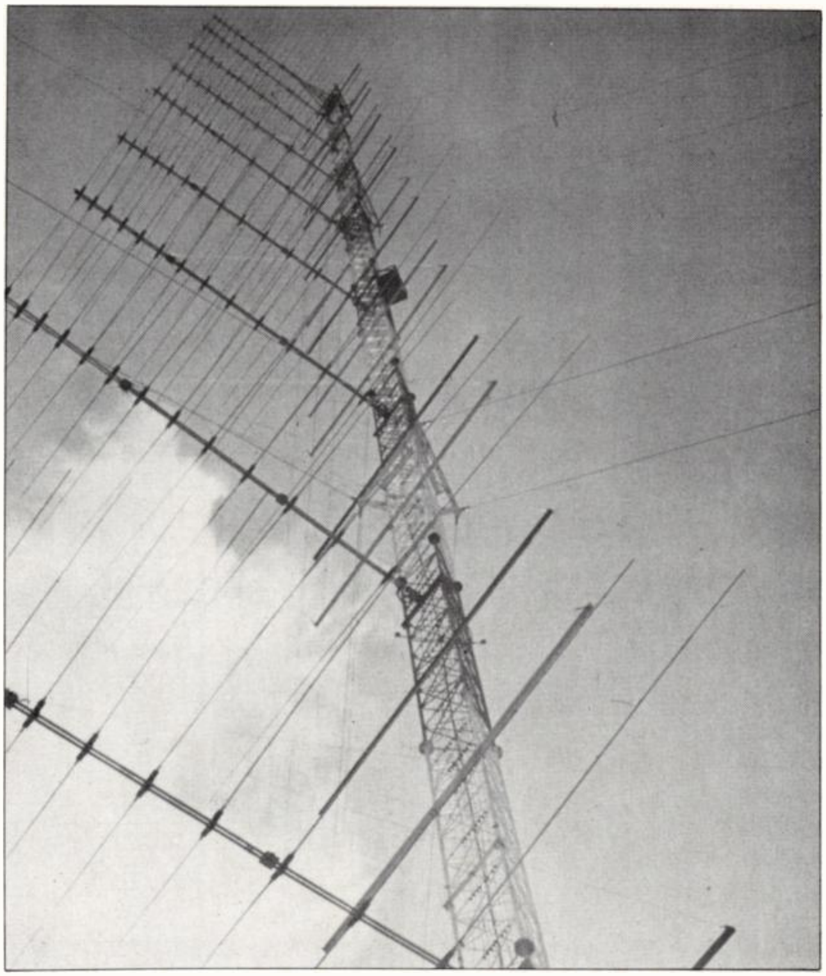

Photo of log periodic stacked array used in Artificial Radio Aurora experiments courtesy of Mr. Russell B. Stoner, N.T.I.A., Boulder, Colo.

Part II

Antennas In The RF Range

. . . 10 kHz to 2000 MHz. The discussion of RF antennas concludes with Driven, Vertical, Leaky-Wave and Reflector types considered.

Driven Antennas

A driven antenna can be defined as an array of two or more elements, separated in phase (electrical and/or spatial), hard connected via conductor or transmission line(s) and combined so as to develop a specific azimuthal or elevation pattern.

Driven antennas can be "loosely" subdivided into *Broadside* and *Endfire* categories. "Loosely" is meant to imply that some Broadside antennas when spatially rotated 90° become Endfire arrays and vice versa.

- **Broadside**

A broadside configuration is characterized by radiation (or reception) perpendicular, or steerable from the normal or perpendicular to the plane of the antenna or array. Four distinct types of Broadside arrays are considered though it must be reiterated that certain types of Endfire arrays can by spatial reorientation be considered Broadside.

The four types are:
1. Curtains
2. Dipoles, Collinears
3. Trombone T
4. Lazy H

Curtains

a. General

A curtain consists of an array of identical radiating elements configured in the vertical plane and in some cases backed by a reflecting screen.

Narrow beams and high broadside directivity gains are achieved by controlled amplitude and phasing techniques. Linear phase variations across the array enable the steering of transmitted or received wavefronts.

A uniformly spaced, uniform amplitude array has the advantage of producing the narrowest azimuthal beamwidth.

A uniformly spaced, tapered amplitude array such as a Binomial or Dolph-Tchebyschev distribution is utilized to minimize sidelobe level — a very useful technique in reducing susceptibility of a receiving antenna to interference.

b. Specific Designs

Each of the curtains illustrated below have substantial gain, various feedpoint locations and are inexpensive to build.

(1.) Sterba Curtain (Figure 15)

Probably the best known broadside array made up of a combination of parallel and collinear elements is the Sterba curtain.

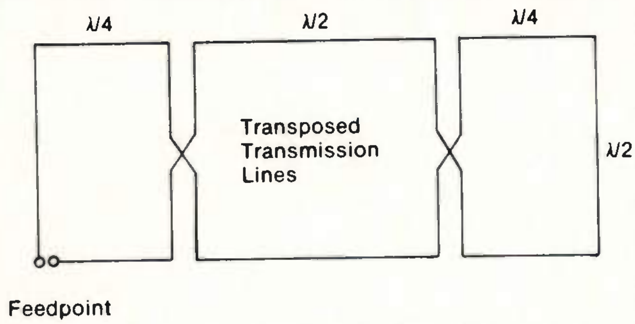

Figure 15. Sterba Curtain

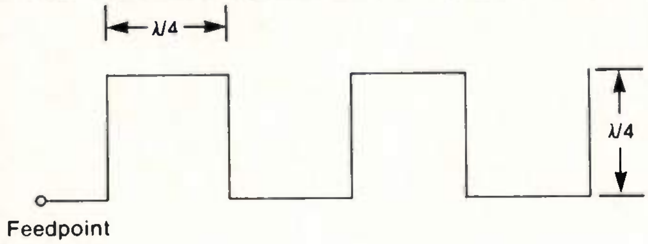

Figure 16. 5 Element Bruce Array

(2.) 5 Element Bruce Array (Figure 16)

The Bruce Array is configured with alternating λ/4 long vertical and horizontal elements. The radiation from the vertical sections are in phase and add while the radiation from the horizontal sections are out of phase and cancel. Typical gain of the 5 element array is 4 dB above a single quarter-wave vertical element.

(3.) The Chireix-Mesny Curtain

The C-M curtain consists of λ/2 dipoles configured in the form of diamonds. It is a center fed array with a broadside vertically-polarized radiation pattern. (Figure 17)

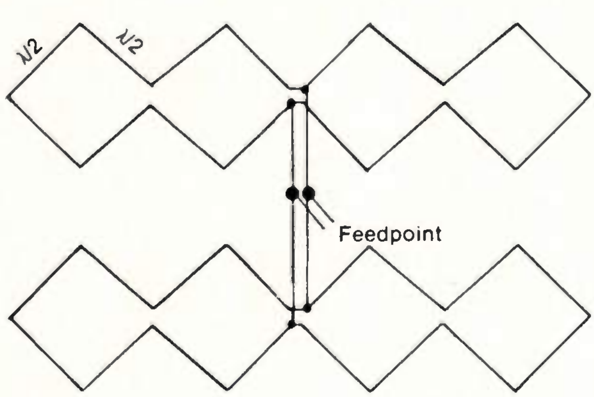

Figure 17. Chireix-Mesny Curtain

(4.) Bobtail Curtain

The Bobtail Curtain is a broadside array of three top current fed quarter wave verticals. Each vertical end section is separated from the middle section by 180°. By virtue of only feeding the center vertical a 1:2:1 amplitude distribution occurs with its associated negligible sidelobe pattern. Theoretical gain is 4.9 dB over a single λ/4 vertical (Figure 18)

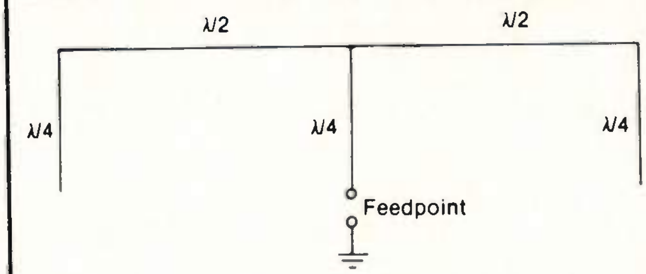

Figure 18. Bobtail Curtain

(5.) X-Array

The capital X array as indicated in Figure 19 has a power gain of 7 dBd.

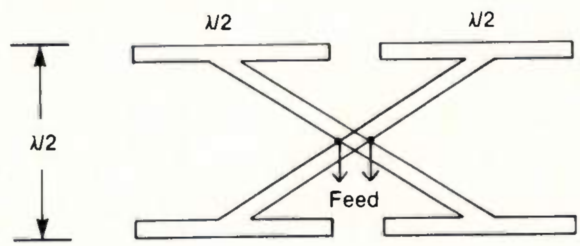

Figure 19. X-Array

Dipoles and Collinears

a. Unstubbed

There are two distinct categories of unstubbed dipoles: center and off-center fed.

(1.) Center-fed Dipoles

Flat-tops and inverted-vees. Both configurations and important properties are tabulated below by length:

Length	Name	Gain
λ/2	half-wave	0 dBd*
λ	double Zepp	1.8 dBd
1.28λ	extended double Zepp	3.0 dBd

*(This is the reference dipole)

The extended double Zepp is the longest center-fed dipole with a radiation pattern remaining at right angles to the line of the antenna. (Max. energy).

(2.) Off-center Fed

Antennas and their characteristics are tabulated below:

Antenna	Figure	Characteristics
Zepp	20-A	Voltage fed with balanced transmission line
Windom	20-B	Resonates on even order harmonics
Fuchs	20-C	Long wire brought directly to transmitter

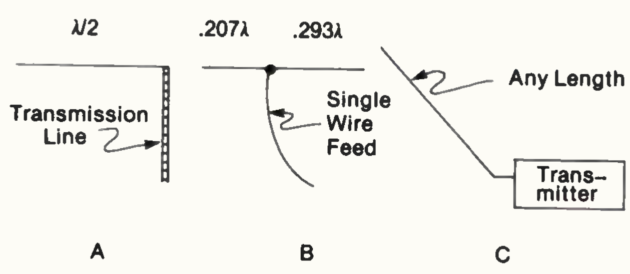

Figure 20. Off-Center Fed Antennas

b. Stubbed

Stubbed "dipoles" are in-phase radiating elements arranged in a single line or collinear. The stub between elements insures that each successive element is fed at the same phase relationship.

A three element horizontal collinear is illustrated in Figure 21.

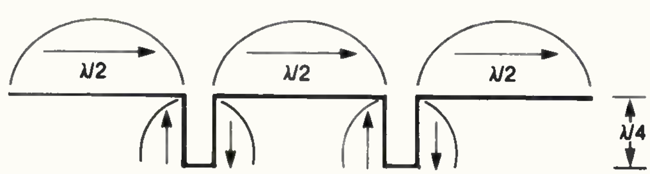

Figure 21. 3 Element Horizontal Collinear

It is a simple antenna and can be fed at almost any point. Additional gain is achieved simply by adding another stub and half-wave element.

The vertical counterpart of this collinear is called a Marconi-Franklin array. (Figure 22)

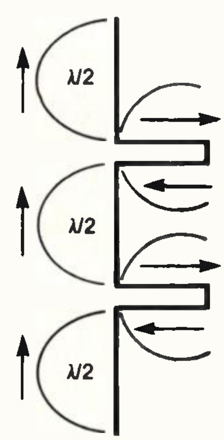

Figure 22. 3 Element Marconi-Franklin Collinear

In both antennas current direction is indicated by the arrows.

3. Trombone T

The Trombone-T is a simple 4 dBd gain broadside antenna of-low-cost. It is illustrated in Figure 23.

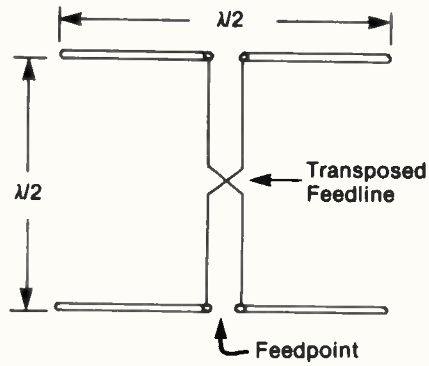

Figure 23. Trombone T

4. Lazy H

The Lazy-H is a twice as large version of the Trombone T. Gain is approximately 6 dBd (5.8 dBd) and it is voltage instead of current fed. See Figure 24.

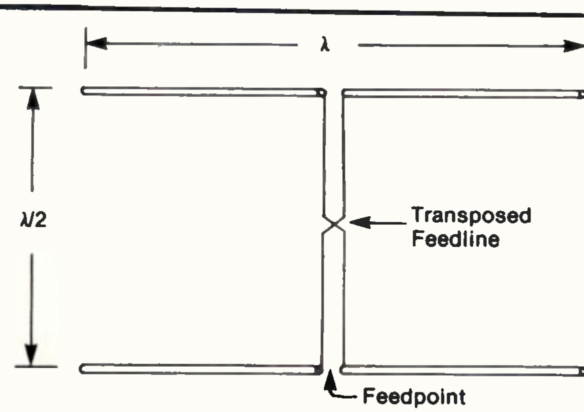

Figure 24. Lazy H

• Endfire

An endfire configuration is characterized by radiation in the same plane formed by the antenna or array. It can be further subdivided into the broad categories of Frequency Dependent or Frequency Independent.

• Frequency Dependent

Antennas that operate over a narrow band i.e. maintain directional patterns and feedpoint impedance response appropriate for its intended use are considered in the classification — Frequency Dependent. They are:

1. 8 JK
2. Double Triplex Beam
3. ZL Special
4. Cardioid Beam Antenna

1. 8JK

The 8JK was named after its inventor John D. Kraus W8JK (Radio amateur call sign). It is an endfire array with approximately 4.3 dBd bidirectional gain. It is a simple antenna to construct with reliable performance. It is illustrated in Figure 25.

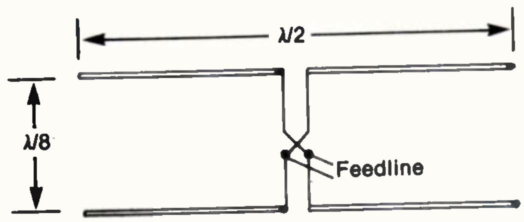

Figure 25. 8JK Antenna

2. Double Triplex Beam

The Double Triplex Beam is basically an 8JK with the single linear conductors replaced by triple leads. It has the added advantages over the single 8 JK of:

— Broadband frequency response
— High radiating efficiency (300 ohm feedpoint impedance)
— Good wet weather performance. (See Figure 26.)

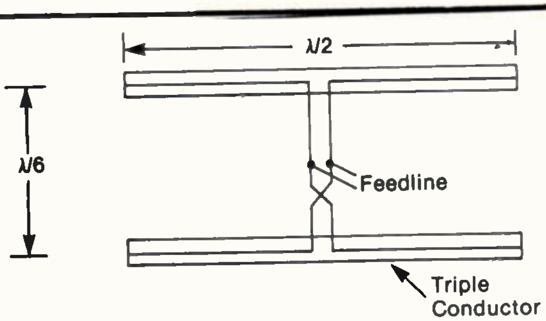

Figure 26. Double Triplex Beam

3. ZL Special

The ZL Special is a two element driven endfire array. The elements are electrically driven 135° out of phase and are spatially phase separated by an additional 45°. It is unidirectional, has a 20 dB front to back ratio and is lightweight and easy to construct. (Figure 27)

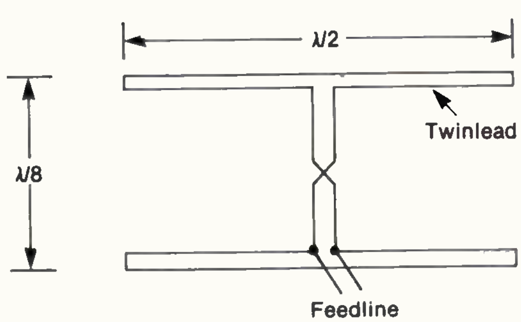

Figure 27. ZL Special

4. Cardioid Beam Antenna

A unidirectional endfire pattern is generated if the total-phase separation (electrical & spatial) between elements is 180°. (i.e. the respectively generated wavefronts must combine in 1 direction (0, 360°) and cancel in the opposite direction (180°). The Cardioid Beam Antenna develops this 180° phase separation with a 90° electrical and 90° spatial phase condition. The resulting pattern is of a broad-nosed cardioid response. (Figure 28)

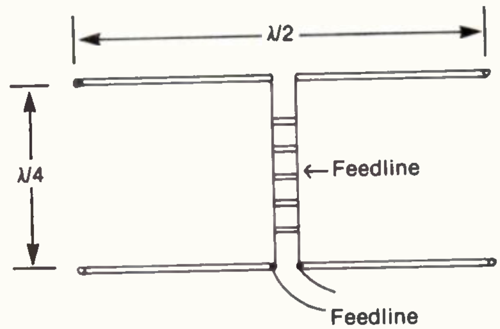

Figure 28. Cardioid Beam Antenna

- **Frequency Independent**

Conversely, antennas that maintain a relatively constant terminal (feedpoint) impedance and vertical angle (elevation angle) or display other frequency invariant directional properties, over several octaves, are classified as frequency independent. The logarithmically periodic array or log periodic for short is a prime example of an RF range frequency independent antenna.

Log Periodic Arrays[36]

A Log Periodic array is a multi-element end-fire driven antenna in which adjacent elements length and spacing are equal to a constant ratio τ.

A second parameter σ is related to the ratio of the separation between adjacent elements and 4 times the length of the shorter element. An illustration and set of formulas will help clarify. See Figure 29. (On page 15).

The gain of a Log Periodic is a function of τ and σ. Other design criteria are:

$$2\ell_1 = 0.47\,\lambda_{max}$$
$$2\ell_{min} \leqslant 0.38\,\lambda_{min}$$
$$R_1 \cdot R_{min} \geqslant 0.50\,\lambda_{max}$$

With these criteria and the required high and low frequencies 3 nomographs are used to complete the design.

Verticals

The simplest of vertical antennas is sometimes represented as the evolution of an open-ended

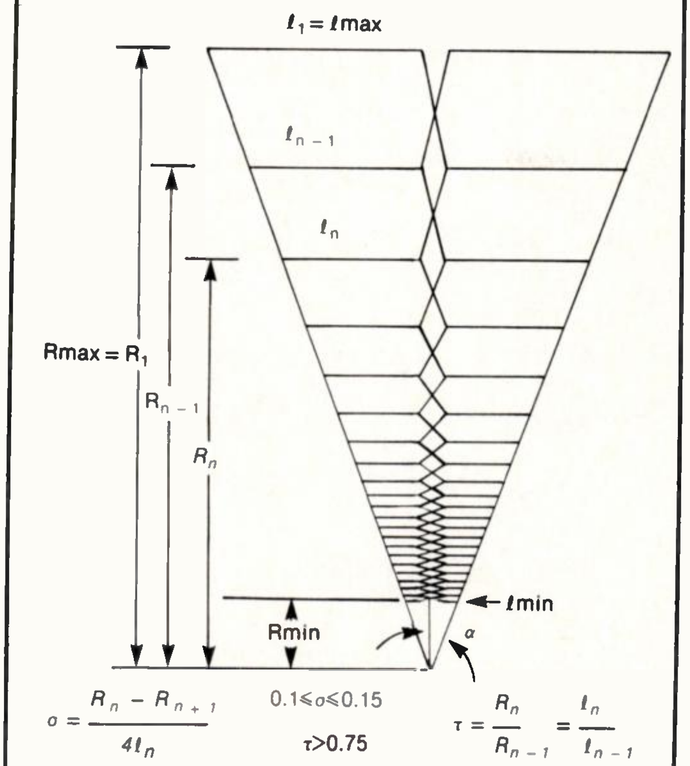

Figure 29. Log Periodic

coaxial transmission line. The outer conductor is folded back exposing the inner conductor and the combination of inner and outer conductors form a complete radiating system or antenna with specific feedpoint impedance and directional pattern.

The vertical antenna is probably the most widely recognized radiator by the public due to its prolific utilization by commercial AM broadcast stations, naval whips and hobbyists in the form of amateur and CB car and roof-mounted ground planes.

Discounting re-oriented horizontal antennas and special categories such as the stubbed collinear or Franklin array, discussed earlier, verticals can be classified as follows:

1. Single driven conductor, length $\leqslant 5/8\ \lambda$
2. Single driven conductor, length $> 5/8\ \lambda$
3. Broadbanded verticals
4. Special application verticals

1. Single Driven Conductor, Length $\leqslant 5/8\lambda$

Vertical antennas are commonly utilized at the following heights: 1/4 λ, 1/2 λ and 5/8 λ. All three heights lay down a maximum signal strength (theoretically) at the horizon or zero elevation angle. This requires that the ground mounted vertical work against an infinite extent, infinite conductivity ground system.

The field strength, under these conditions, along the horizontal as a function of antenna height for a vertical grounded radiator with one kilowatt radiated power is tabulated below:

Antenna Height	Field Strength
λ/4	196 mv/meter
λ/2	236 mv/meter
5/8 λ	276 mv/meter

Such is rarely the case. Usually, in commercial service, (AM broadcast) 120 radials are laid out every 3 degrees. This does help stabilize the impedance of the vertical system though in itself does *not* lower the maximum radiation angle to the horizon.

Simple techniques are however available for establishing an extremely low (as low as 0°) elevation or takeoff angle. Mounting a vertical above ground with a radial or counterpoise system drooping downward at about 45° helps to significantly lower the elevation angle (of maximum radiation). One precaution worthwhile mentioning is that the counterpoise should be at least a λ/4 above ground.

Numerous studies have developed working relationships between vertical antenna efficiency, length and number of radials. [26,27,28,29]

2. Single Driven Conductor, Length >5/8λ

Verticals greater than 5/8λ in continuous length are difficult to construct for the lower frequencies and as a result of vertical angle lobe splitting and its concomitant high angle radiation are not as widely utilized as the previous category.

3. Broadbanded Verticals

There are several vertical antennas that operate over large bandwidths with VSWRs not exceeding 2.0:1. Two of them, the Discone and the Conical Monopole are discussed.

a. Discone

The discone is a wideband vertical radiator that can be directly matched to a 50 ohm system. The basic configuration of a Discone is indicated in Figure 30.

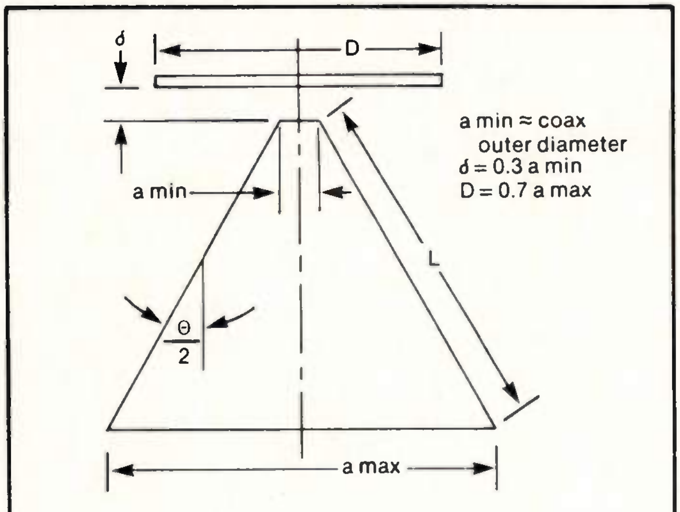

Figure 30. Discone

The VSWR frequency variation at the low end of the band is determined by L and θ. An optimized discone with $\theta = 60°$ and $a/L = 1/22$ has a VSWR of less than 1.5:1 over a 7 to 1 frequency range and a VSWR of less than 2.0:1 over a 9 to 1 frequency range.[30]

b. Conical Monopole

The conical monopole is short vertical radiator ressembling two terminated cones placed base to base. It operates over a 4:1 bandwidth, and has high radiation efficiency. The configuration with dimensioning is indicated below.[31] (See Figure 31)

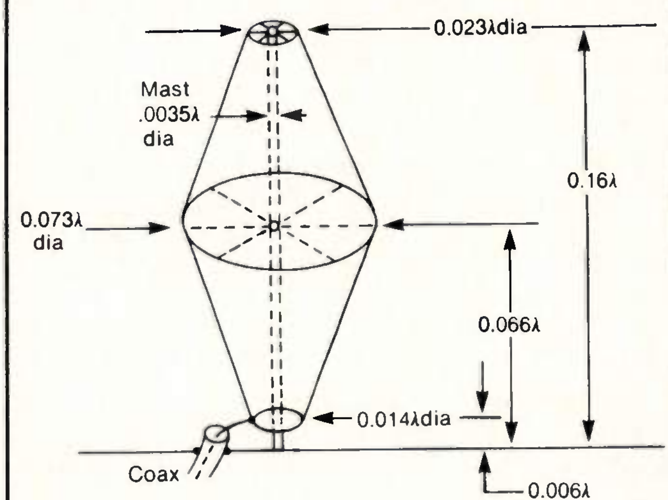

Figure 31. Conical Monopole

4. Special Application Verticals

LF and VLF applications are probably the hardest areas to accommodate in terms of antenna design due to several conflicting constraints. Foremost is the operating wavelength — in a word — long. Reasonable feedpoint impedances, bandwidth and

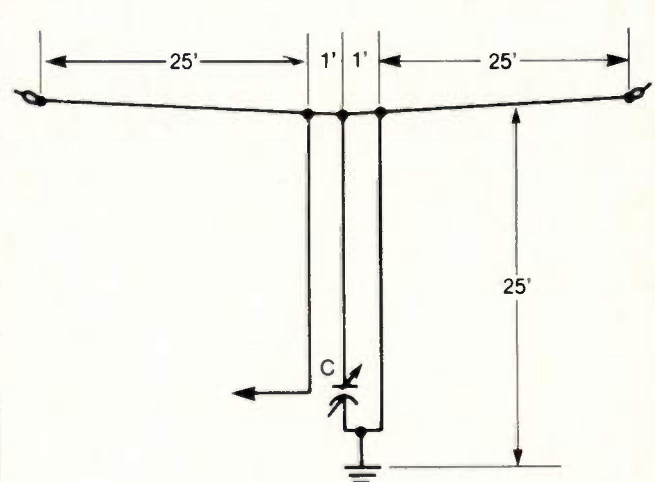

Figure 32. UG Antenna

efficiencies must be maintained. Two, out of many configurations, that are in themselves tradeoffs, are discussed.

a. Type UG

The type UG antenna has the shape of a capital T as illustrated below with dimensioning given for 2 MHz operation. Capacitor C is varied until a favorable feedpoint impedance is developed.[32] (Figure 32)

b. NORD

The NORD is a short vertical antenna, used at LF and MF. It has a sufficiently high bandwidth and radiation efficiency for its intended use. It is basically three over-coupled tuned circuits. A center tower is used as a common element. See Figure 33.

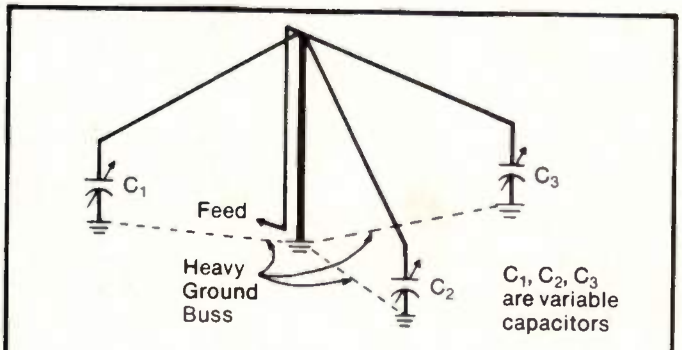

Figure 33. Nord Antenna

Leaky Wave Antennas

Leaky wave antennas can be identified by their low silhouette (close to ground) or controllable aperture distribution i.e. beamwidth and sidelobe level. Several types identified here are:

1. DDRR
2. Fast Wave
3. Slow Wave

1. DDRR

a. λ/4 DDRR

The Directional Discontinuity Ring Radiator developed by J.M. Boyer is an omni-directional low profile vertical polarization radiator. It is a single circular open-ended element mounted horizontally a short distance above a ground plane. See Figure 34.

According to theory the DDRR antenna ring forms a transmission line with its ground plane. The slot, between the ring and ground plane being a discontinuity, radiates continuously in a direction transverse to the ring axis.

The advantages of this antenna include small horizontal dimensions, extremely low height and simple construction. Since it is DC grounded induced charge static is not a problem as it is with a monopole.

b. λ/2 DDRR

A larger version of the original DDRR is available. Primary advantage is an increase in gain. It is as depicted in Figure 35.

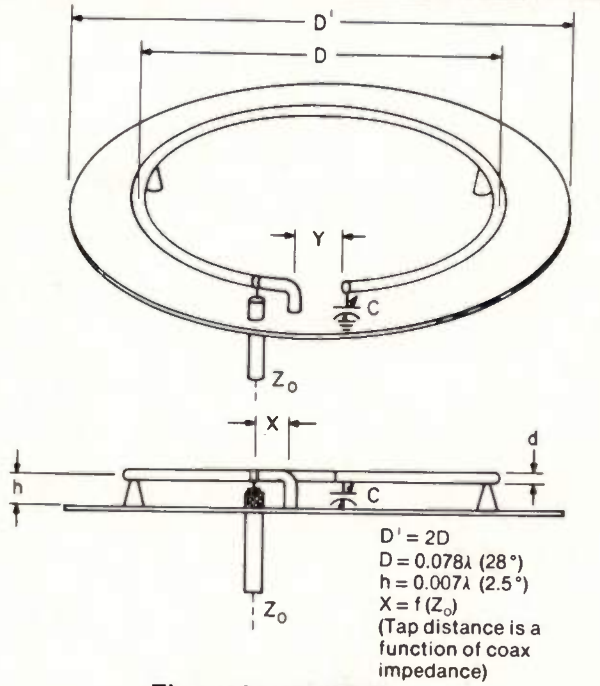

Figure 34. λ/4 DDRR

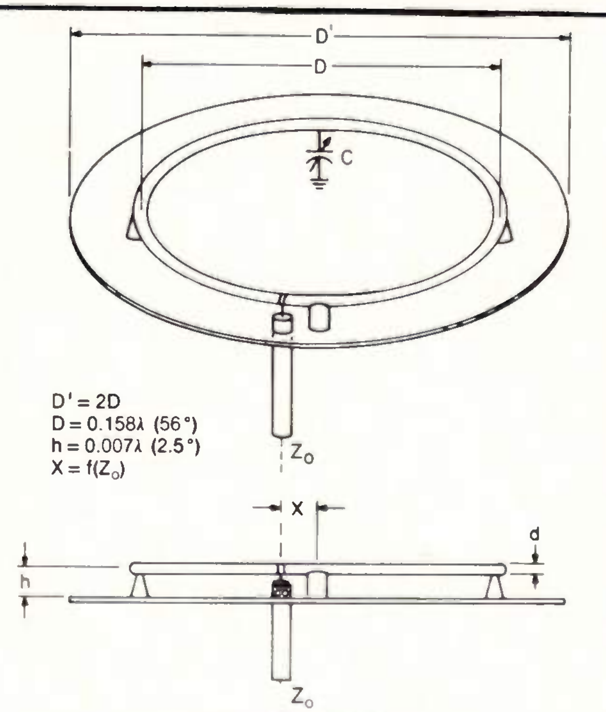

Figure 35. λ/2 DDRR

2. Fast Wave

Another class of leaky wave antenna is the Fast Wave Structure. Similar to the DDRR it is a low silhouette structure. It is cut along its length to permit the radiation of energy[34]. The angle of radiation is closely related to the phase velocity of the guiding structure before the cut is made. The gaps must be smaller than λ/3 of the highest frequency of operation.

3. Slow Wave

An additional category of leaky wave antenna is a slow-wave structure. It is periodically-loaded (series capacitors or shunt inductors) long wire antenna. The radiation angle is determined by the phase velocity of the slow wave and the periodicity of the loading. The spacing between successive loadings must be approximately a half-wavelengh or greater, for with closer spacings no radiation occurs.

Reflector Antennas

Beam formation in this type of antenna relies on the shape of the reflecting surface to focus energy from a line or point at the feed position into a narrow parallel wavefront beam. The converse also holds i.e. intercepted parallel wavefronts are focused into a common line or point location. Two types considered are:

1. Corner Reflector

A corner reflector is a half-wave radiator with a reflector consisting of two plane metal surfaces meeting at an angle immediately behind the radiator.

The radiation is the sum of direct radiation from the dipole and reflections from the planes. Consequently the pattern depends on the distance from the dipole to the corner.

At VHF the corner reflector is often solid (metal). At lower frequencies screen wire or metal rod constructed reflectors are used. The maximum gap between the wires or rods should not exceed 0.06λ.

Power gains of at least 10 dB to as much as 14 dB are achievable using a corner reflector. A diagram with design information is shown in Figure 36.

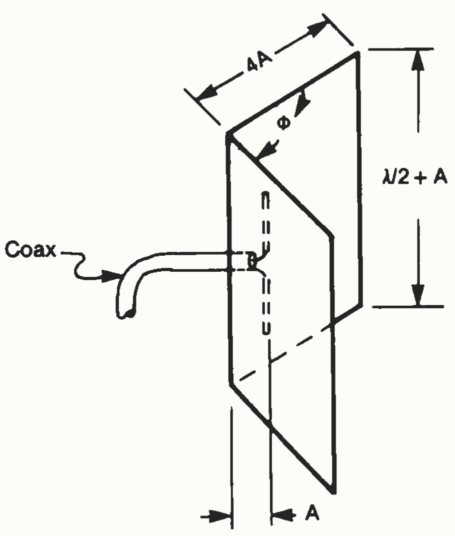

Figure 36. Corner Reflector

Principal advantages of the corner reflector are its simple construction and the low side and back lobes of the directivity pattern.

2. Parabolic Reflectors

Parabolic reflectors are widely used in UHF and microwave radar and point-to-point communications systems to achieve antenna gains in excess of 20 dB.

Recently the CATV industry has been successfully utilizing a Half-bolic to receive very weak over the horizon VHF signals.[35] (A Half-bolic, as its name implies is physically half a parabolic reflector. When combined with its ground image its performance is similar to a parabolic reflector.)

In the low frequency application an offset feed is employed to reduce undesired interaction between reflector and feed and to minimize aperture blocking.

A suitable feed might consist of a log periodic array in which the longest element forms the dipole feed of a corner reflector. The phase center of the corner reflector would be located at the focus of the parabola. Using this technique the maximum possible gain and narrowest possible beamwidth are achieved at the lowest operating frequency.

Though 5 types of antennas with 51 different configurations were discussed this is not to imply that these are, or were the only antennas ever designed. This is a broad sample of most common antenna configurations utilized by all services — commercial, government and hobbyist. Modifications such as shortening, employing active elements and material experimentation affects the properties and probably the classifications of the antennas already discussed.

Future articles will delve into specialized antenna configurations utilizing some of the above-mentioned techniques, and others not yet considered. □

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Scalar Network Analyzers

By r.f. design *Staff*

The life of an RF engineer 15 years ago as historically described in the editorial was simpler without the 1000 MHz video bandwidth scopes and 11 digit precision frequency counters. And yet it's almost painful to remember the number of hours spent on the lab bench **not** on the design or fabrication of RF circuits but on its testing. That has changed for the better with the availability of among other items a Scalar Network Analyzer.

A Scalar Network Analyzer is a system that helps measure the scalar or amplitude response of "devices" be it insertion loss (transmission loss or gain), return loss or output power levels. The "devices" can be filters, attenuators, amplifiers or any other two port device. It is effected by sine wave testing over the frequency range of interest.

Historically early scalar measurements were made at fixed frequencies in a point-to-point fashion. These measurements were tedious, time consuming, and often yielded incomplete results. For instance, resonances between points of measurement were often missed or the skirt responses of filters and amplifiers were inadequately defined.

The evolutionary process of overcoming these difficulties was initiated in 1952 with the introduction of the first broadband, high directivity directional coupler. The ability to separate forward and reverse traveling waves on a broadband basis speeded fixed-frequency reflection and transmission measurements. In 1954 swept-frequency measurements became possible with the introduction of a sweeping signal source using a motor-driven, mechanically-tuned klystron. These instruments combined with ratio-meters and broadband detectors to introduce swept measurements as a powerful new methodology for network optimization.

By 1957 the mechanically-tuned klystron had been obsoleted by the voltage-tuned backward wave oscillator tubes (BWO) as a source of swept RF signals. The voltage-tuned BWO's allowed faster and more accurate sweep times while relieving the problems of moding and tracking associated with klystrons.

Solid-state microwave sweep oscillators were developed in the late 1960's to overcome the finite life problems associated with BWO's. Further development led to the introduction of multi-octave solid-state sources, allowing continuous swept measurements over several octaves of frequency.

Innovations in other areas of swept-measurement technology complemented the development of sweep oscillators. Multi-octave, high-directivity, coax-directional couplers were developed at frequencies up to 18 GHz. New diode detectors with more sensitivity were combined with the appropriate ratiometer display to obtain greater measurement range and increased measurement confidence over broad frequency spans.

Also in the late 60's, several companies introduced a calibrated memory audio substitution data normalizer system using an 8 track cassette tape as the storage medium. It wasn't until 1974 when the first high speed digital memory, which provided a flicker-free display, was introduced.[2] 1975 saw the publication and broad-based acceptance by the Electronics Industry of IEEE Standard 488-1975[3]. This standard enabled instruments to "talk" and "listen" to each other directly or under the auspices of a (computing) controller.

Scalar Network Analyzers regardless of the size, cost or degree of automation and whether actually called by such nomenclature or simply described as a **Microwave Multimeter**[4] all have several common elements. They are:

1. Swept Frequency Sources
2. Signal Separation Devices
3. Detectors
4. Displays (and Processing)
5. Connectors and Cables

There are many U.S. manufacturers of Scalar Network Analyzers (detailed in this article) each with their own setups and procedures but a typical configuration for insertion loss and return loss can be illustrated **generally** as in Figures 1 and 2.

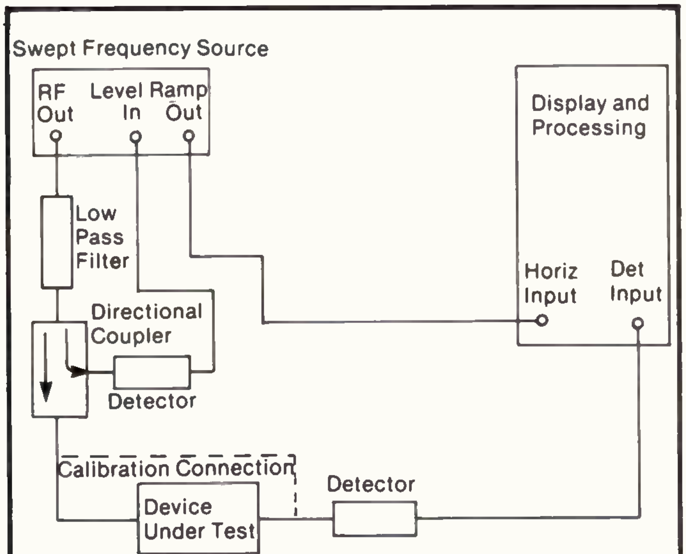


Figure 1. Insertion Loss Test Set-up.

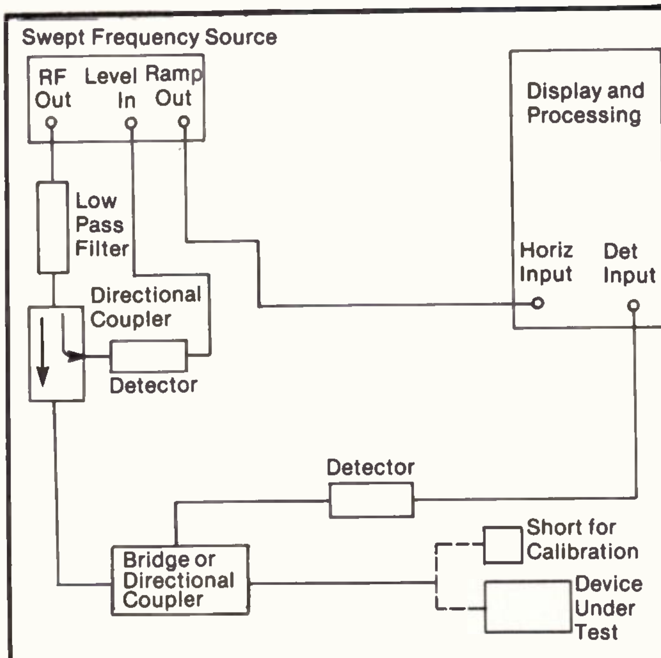


Figure 2. Return Loss Test Setup.

Insertion Loss Measurements

After the preliminary steps (set-up and frequency source/display settings) have been accomplished a calibration procedure is started. This consists of "zeroing out" inherent system residuals by connecting the detector directly to the coupler as shown in Figure 1 (dotted line). Store and subtraction modes are entered and we are ready to begin testing the D.U.T. (device under test). In effect digital circuitry has replaced (more accurately and rapidly) the old grease pencil that was used to establish our "zero reference" line or condition on a scope trace. The "best-looking" display of the swept frequency response of the D.U.T. is arrived at through various "OFFSET" and "N dB/DIV" display control adjustments. Data is displayed (transmission gain/loss swept response), read, printed, plotted or otherwise manipulated as per the tester's needs and peripherals. ("Store and subtraction" is further discussed under "Automation".)

Return Loss Measurements

The procedure is similar to Insertion Loss measurements. The differences occur depending upon the calibration method chosen. Figure 2 illustrates a set-up and calibration technique. Here a short is placed on the bridge or directional coupler output and the response stored in memory.

Another procedure utilizes two directional couplers joined together to form a reflectometer. During calibration the reflected wave coupler is turned around so that it is sensitive to the incident wave. The short is placed and data taken. The coupler is returned to its original placement and the D.U.T. is installed. The insertion loss of the coupler (during calibration) is allowed for by adjustment of offset controls on the display and processing unit. Handling of data is as before.

Power Level Measurements

Absolute power measurement techniques vary depending upon which manufacturers analyzer is utilized. The first method considered[5] requires that HIGH and NOISE level calibration adjustments first be made. The D.U.T. is swept and the display trace is examined at the frequency of interest.

OFFSET thumbwheel switches are used to "null" the unknown (level) to the display centerline. The absolute power level at **that** frequency is then read from the thumbwheel settings ± the detector (frequency response) calibration correction. Several other analyzers[6,7] utilize the same basic procedures with differences mainly in nomenclature (control settings).

A multimeter type of network analyzer[4] measures absolute power level by following a user prompt sequence of instructions indicated on a liquid crystal display. A calibration routine is automatically initiated (if this is a first of a kind measurement) followed by the connection of the D.U.T. The correct calibration factor is generated upon operator selection of desired frequency. All zero-bias Schottky diode detectors supplied by this manufacturer comes replete with interchangeable calibration ICs for inputting the linearization correction microprocessor internal "look-up" tables. (This sure helps if you happen to accidentally drop and break the original detector.)

If it were only for the **Insertion Loss, Return Loss** and **Power Level** measurement capability the Scalar Network Analyzer would still hold its own as a versatile and needed piece of test equipment for the RF design engineer. However, with a little bit of ingenuity other applications are possible that greatly enhance the analyzer's utilization as a design tool. A few of these applications are discussed.

1 dB Gain Compression

The input level at which an amplifier's small signal gain decreases by exactly 1 dB is a universally accepted criterion used by design engineers in specifying a limiting condition of linear device performance. With the Scalar Network Analyzer set-up, as shown in Figure 3A, a CW gain compression measurement is accomplished as in Figure 3B.

The horizontal sweep ramp is used to modulate the RF input and to provide CRT horizontal deflection proportional to power while the vertical channel is used to indicate the difference between the input and output levels.

1 dB Gain Compression Frequency

It is possible to determine the **frequency** at which 1 dB gain compression occurs in an amplifier with a slightly different (and simpler) set-up and procedure. The amplifier is swept and its small signal gain versus frequency is stored in memory. The input level is increased until input minus the stored response differs by 1 dB.

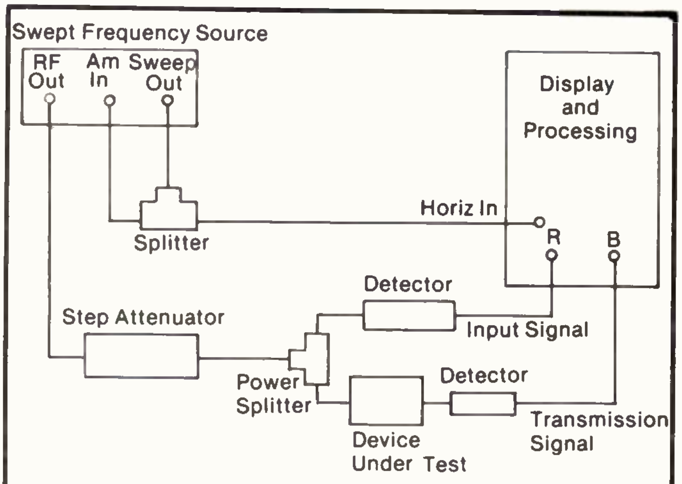


Figure 3A. CW Gain Compression Set-up.

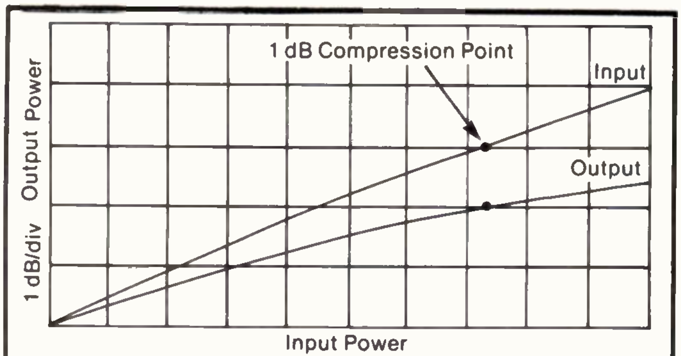


Figure 3B. 1 dB Gain Compression Measurement.

Comparison Testing

A natural application of a Scalar Network Analyzer is to compare the swept response of an unknown device against a standard unit. The "unknown" can be an amplifier, filter, coupler or attenuator in which adjustments are made while sweeping the units. A null between the swept response of the unknown and standard unit (or its stored version) is tuned for or measured against acceptable specifications.

Harmonic Measurement

An interesting and useful display of a devices' performance can be accomplished by sweeping the D.U.T. over its fundamental frequency range. Its output is power split with one leg feeding channel 1 of a display while the second leg is filtered (fundamental eliminated) and displayed on channel 2. Second and higher order harmonics are indicated and can be referenced in power level against the fundamental actual or stored swept response.

Component Specifications

As mentioned earlier there are five integral "components" that make up a Scalar Network Analyzer. Each one provides capabilities that are described below.

Swept Frequency Sources

Sweep Oscillators (sweepers), a source of swept RF signals, available over a 6 decade frequency range (100 kHz-110 GHz)[8] must meet minimal requirements to be a useful adjunct to Scalar Network Analyzers.

Sufficient Output Power, which can be leveled, is needed for wide dynamic range performance. Output levels of 10 milliwatts (+10 dBm) are typical. Associated with sufficient level is flatness of the output response over the entire frequency range of interest. Though arguments can be made for internal and external leveling techniques as well as storage subtraction of system response (to be discussed) and ratio procedures an **inherent** flat-frequency sweeper response is desirable.

Source Match or the lowest available VSWR looking into the sweeper (lowest return loss) over the entire frequency range reduces a major contribution of measurement error. An available specification for an externally modulated sweeper over the 2 decade range of 40 MHz to 4 GHz has a return loss equal to or better than 15 dB.[9]

Accurate Linear Sweeps and calibrated frequency markers are important for frequency determination and horizontal calibration of CRTs and X-Y recorders.

Frequency Stability is necessary for making accurate single-frequency and narrowband measurements.

Low Harmonic Content is required to minimize measurement inaccuracies when broadband detection is used.

Coverage of the entire frequency range of interest **without interruption** has always been an essential operations criterion even prior to Network Analyzers. It reduces the possibility of introducing errors possible as a result of plug-in head changes.

Signal Separation Devices

The three most commonly used devices for signal separation in high-frequency swept measurements are:

- Directional Coupler
- Directional Bridge
- Power Splitter.

Directional couplers are devices used to separate or sample the traveling wave moving in one direction on a transmission line while remaining virtually unaffected by the traveling wave moving in the opposite direction.

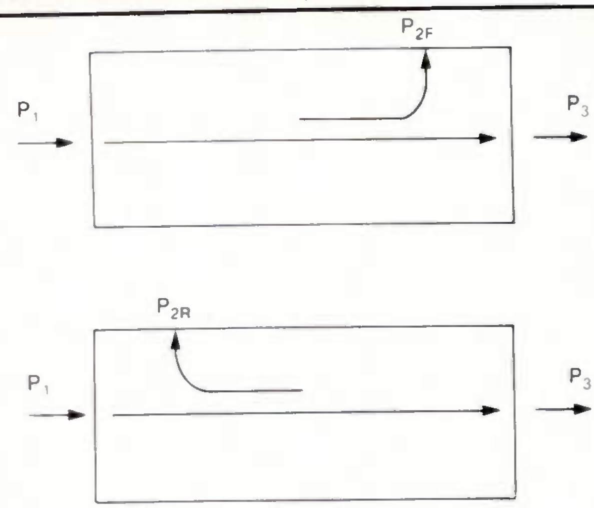
Coupling factor, directivity, mainline loss, and mainline SWR are the four principal parameters defining directional coupler performance. **Coupling factor** is the fraction (in dB) of the power moving in the forward direction through the coupler main arm that appears at the secondary arm.

Since directional couplers are not perfectly directional devices, **directivity** is used as a figure of merit for a coupler's ability to separate the forward and reverse traveling waves. **Directivity** is the ratio (in dB) of the power in the secondary arm when all the power is flowing in the forward direction of the main arm to the power in the secondary arm when an equal power is flowing in the reverse direction of the main arm.

The **mainline loss** is the fraction of power lost by a traveling wave moving in either direction through the main arm of a directional coupler.

Mainline SWR is a measure of mismatch at the coupler output port and is important in qualifying effective source match.

These relationships are indicated by the diagrams and formulas in Figure 4.



$$\text{Coupling Factor} = -10 \log \frac{P_{2F}}{P_1}$$

$$\text{Directivity} = -10 \log \frac{P_{2R}}{P_{2F}}$$

$$\text{Mainline Loss} = -10 \log \frac{P_3}{P_1}$$

Figure 4. Directional Coupler Relationships.

Directional bridges utilize impedance bridge techniques to accomplish the same objectives as directional couplers. Main arm to auxiliary arm coupling, directivity, and mainline SWR are the important parameters in qualifying directional bridges. Main arm to auxiliary arm coupling is equivalent to the sum of the mainline loss and coupling factor in a directional coupler.

Power splitters are usually defined by their main arm insertion loss, equivalent output SWR, and output tracking. Main arm insertion loss is the loss that occurs between the input and either of the two output arms.

Output tracking (in dB) indicates how well the splitter maintains equality between P_{out1} and P_{out2}.

Equivalent output SWR is a measure of how good a source match may be obtained when a power splitter is used in leveling a sweeper or in a ratio measurement. Power splitters may be of either two-resistor or three-resistor construction; three-resistor power splitters are greatly inferior to two-resistor power splitters in terms of equivalent output SWR when used in leveling or ratio applications.

These relationships are indicated by the diagram and formulas in Figure 5.

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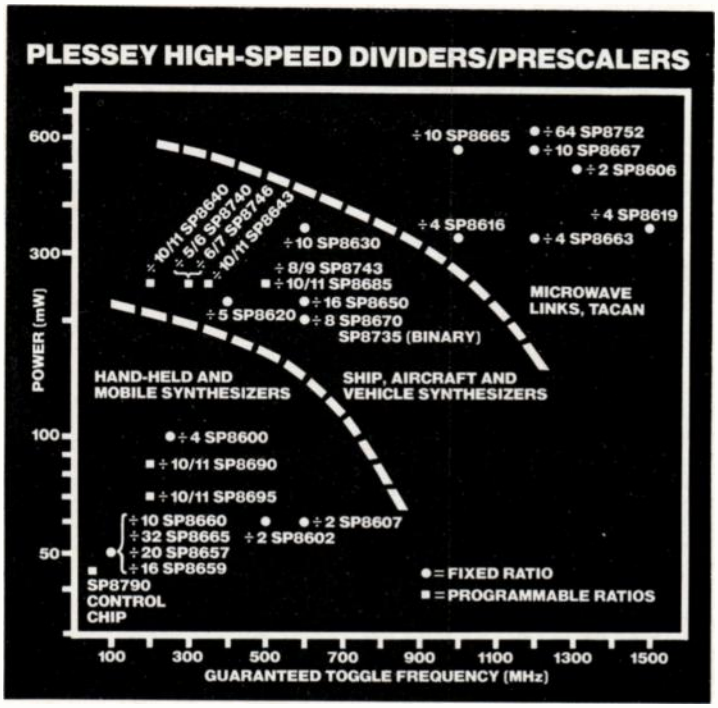

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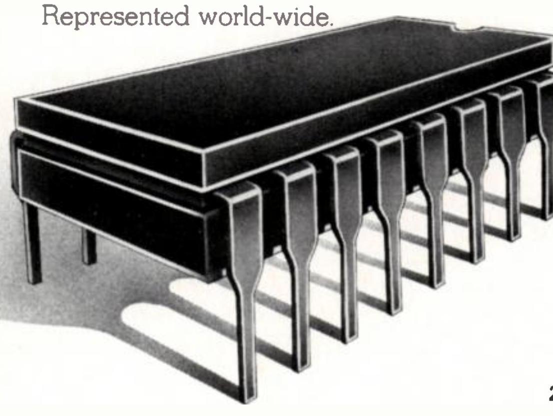

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$$\text{Main arm loss} = -10 \log \frac{P_{out1}}{P_{IN}} \ (6\ \text{dB})$$

$$= -10 \log \frac{P_{out2}}{P_{IN}}$$

$$\text{Output tracking} = 10 \log \frac{P_{out1}}{P_{out2}} \ (<0.2\ \text{dB})$$

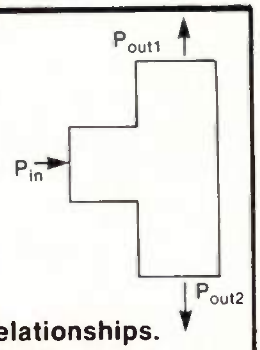


Figure 5. Power Splitter Relationships.

Detectors

Broadband swept measurements require accurate detection of both absolute and relative power levels over wide frequency ranges. In most swept measurements, detection for leveling and display of data is accomplished either by diode detectors or temperature-compensated thermal detectors.

There are five fundamental detector characteristics:

1. Frequency response
2. Reflection coefficient over full band
3. Sensitivity
4. Transfer characteristic
5. Response time.

Diode Detectors

Diodes provide a DC output when an unmodulated RF signal is applied and are widely used in swept microwave measurements because of their simplicity, sensitivity, fast response time, and broad frequency range.

A diode exhibits a square-law transfer characteristic with an output voltage variation that is directly proportional to the applied RF power variation. The two most commonly used diodes in swept measurements are the point-contact (crystal) and the hot-carrier with sensitivities of − 45 and − 50 dBm respectively.

Thermal Detectors

Where accurate knowledge of average absolute power is required, temperature compensated thermal detectors: bolometers and thermocouples are utilized. Thermal detectors sense RF power levels by determining the amount of thermal energy or heat generated in a particular load impedance.

Bolometer detection is based on the fact that a thermal element's resistance changes with the RF power dissipated by the element. When such a thermal element is inserted into a bridge, absolute power can be measured through bridge balancing techniques using a differential amplifier.

Thermistors are characterized by their broad frequency range (typically 1 MHz to 40 GHz), measurement range (1 μW to 10 mW), and low-frequency response. Similarly, thermocouples are noted for their 100 kHz to 18 GHz frequency range, low drift, measurement range (0.3 μW to 100 mW), and exceptionally low reflection coefficient (1.28 SWR at 18 GHz). Both types of detectors have long response times in comparison to diode detectors.

Display and Processing

The greatest **apparent** variation in design and measurement techniques between different manufacturers' Scalar Network Analyzers occurs in the display and processing units. Most units include an internal/external (horizontally) triggered scope with plug-in or hard-wired horizontal and vertical heads. Each vertical head provides thumbwheel or pushbutton scale factor and reference level switches. Most include a digital storage medium that allows for the "memorization" of system residuals needed in a normalization or calibration procedure prior to the testing of a D.U.T. These system residuals by means of "input minus memory" or "memory subtract" result in a swept response display that for the most part are free of any frequency-related idiosyncracies inherent in the measurement system itself.

Display Modes

Four display modes of operation described below are indicative of the capabilities available in Scalar Network Analyzers.[7]

Real Time Display Mode. The horizontal sweep is synchronized with the ramp from a sweep generator and may be varied from approximately 10 msec to 100 sec per sweep.

Refresh Display Mode. The ramp from a sweep generator is digitized and stored in a 1024 point memory (512 points for dual trace). To provide a flicker-free display, the stored data is continuously updated at the sweep generator sweep rate while the refreshed display is swept at a 14 msec per sweep rate. Even though the generator may sweep at a slow rate, the processor presents a steady, non-flickering display.

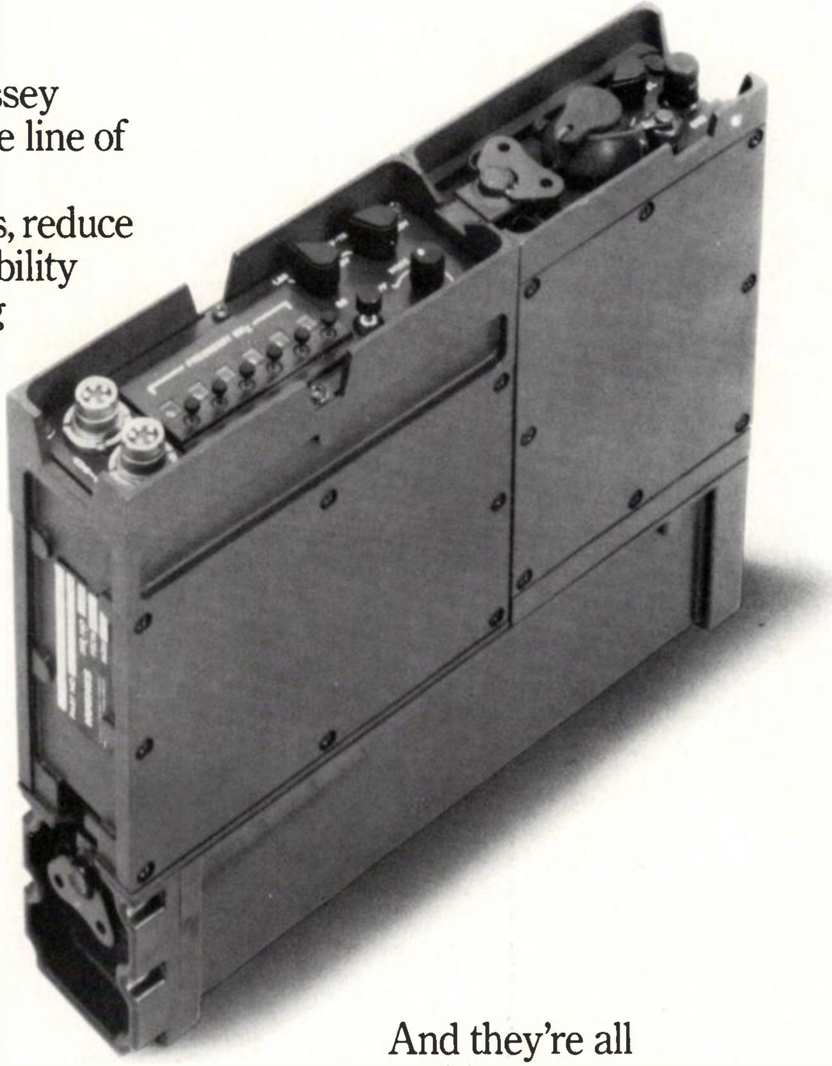
Refresh Hold Display Mode. The updating of the display is stopped, and the display is frozen for analysis or photography.

X-Y Plot Display Mode. The refresh sweep rate is slowed to approximately 30 sec per sweep for making plots on a mechanical recorder. A hard copy of any display, single, or dual trace, may be made. Dual trace recordings are made automatically by sequentially plotting the A and B channels.

Detected Inputs

Detected RF inputs, needed for display and processing, have varying nomenclature though typically are described as "A", "B" and "R". Several manufacturers' units allow for simultaneous presentation of 2 out of 3 inputs. For example simultaneous display of transmission loss/gain and return loss would utilize the "B" and "A" inputs, respectively. The 1 dB CW gain compression set-up, illustrated in Figure 3A, used an input labelled R. This "Ratio" or "Reference" input when used in conjunction with either the A or B

Radio Active



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inputs provides a normalization routine that helps eliminate undesirable externally-varying error-producing signals. An obvious example involves the gain measurement of an amplifier driven by a non-flat (frequency), non-leveled (amplitude) sweeper. By inputting the amplifiers output and sample of its drive signal and forming the ratio, the sweeper's undesirable variations are greatly reduced.

Other inputs founds in most Network Analyzers are briefly described below:

Horizontal Sweep Ramp Input. Maximum of 20V peak-to-peak input signal.

External Blanking Input. +3 to +10V sweeper trigger signal.

Marker Input. 1 mv to 10 volt peak signal, −3 to −10V introduces markers.

Sequential Sync Input. +3 to +10V blanks and maintains trace amplitude during switching of sweep generator oscillators.

Outputs

Horizontal Sweep Ramp Ouput. 0 to 10V in synchronism with sweep display.

Blanking Output. Positive TTL compatible voltage during display retrace.

[**Differences** between manufactured units and their respective specifications are delineated in the "Product Line" section.]

Connectors

Maximum usable frequency response of an analyzer is related to the connectors utilized with the signal separation devices, detectors and, to a lesser extent, the swept frequency source.

Typical 50 ohm connectors utilized by the industry are listed below:

Type N	WSMA
APC-7	SMA
GPC-7	

System Specifications

Scalar Network Analyzer typical measurement system specifications are delineated below:

Frequency Range: 1 MHz to 18 GHz with extension to 110 GHZ

Frequency Response: ±0.5 dB to 12.4 GHz and ±1 dB to 18 GHz

Measurement Range: +10 dBm to −50 dBm

Accuracy: Overall system uncertainty for a single channel measurement is 0.1 dB per 10 dB

Resolution: 10, 5, 1, 0.5 or 0.1 dB per division

Offset: ±99.9 dB in 0.1 dB steps

Temperature Range: +15°C to +45°C operating

Power: 50-60 Hz, 115/230 volts ± 10 percent at approximately 100VA

It is stressed that the above specifications are **typical** and do **not** attempt to represent all manufacturers' specs.

Sources of Errors

The ideal Scalar Network Analyzer would have a perfectly flat (frequency response), constant amplitude, infinite adjustable power, infinite decade coverage, stable, highest purity sweep generator; 1.00:1 VSWR ports on all devices, perfectly linear detectors, etc. The reality of the situation is that none of the above are available and consequently are the causes of errors and uncertainties in this measurement system. However, potential causes of errors can be understood and if not perfectly corrected for at least reduced. The following discussions indicate sources of errors, an analysis and corrective measures.

Source Errors

If the sweep generator output port has **other** than an ∞ dB return loss (and it always does) then the interaction between reflections from the **source match** and the unknown impedance creates a measurement uncertainty. Let us use Figure 6 to follow the signal path.

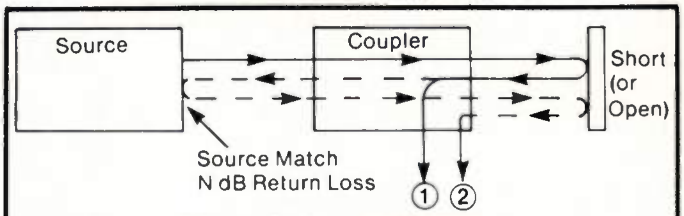


Figure 6. Source Match.

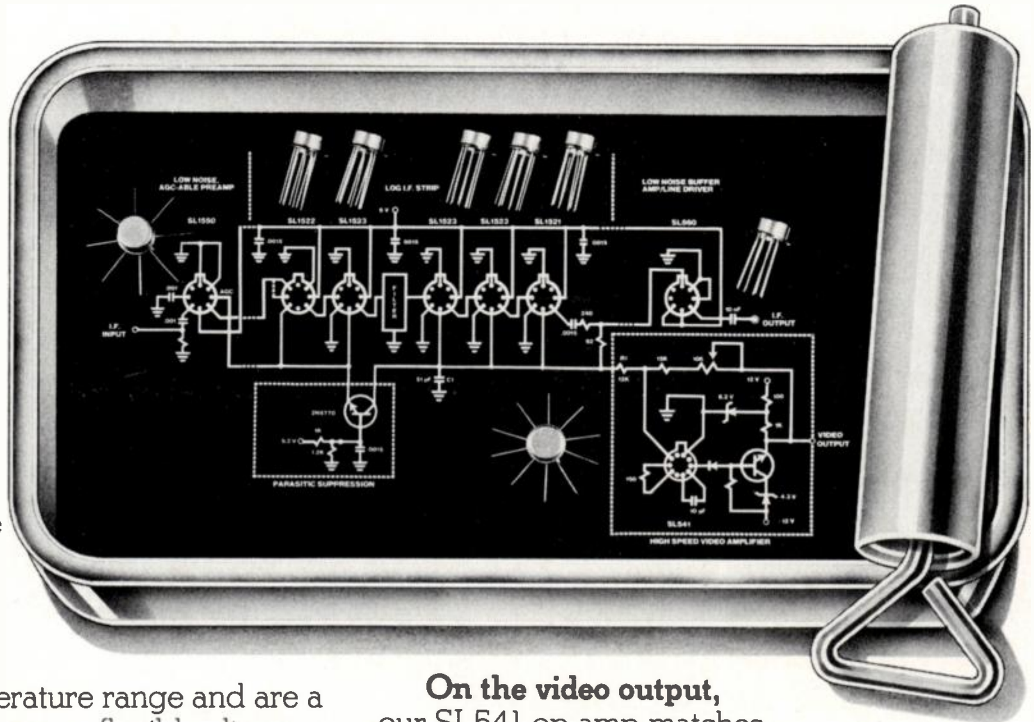
The signal travels through the coupler with little attenuation, experiences a full reflection at the short (or open) and returns to the coupler. A power split occurs, a portion of the signal outputs ① and the majority travels (thru the coupler) back towards the non-perfect source. A reflection occurs (N dB down) which repeats the previous routine and exits at the coupled port ② The difference in signal level between the desired ouput and the re-reflection is slightly greater than N dB, the source return loss. These two signals can be in, out of phase or any combination over the entire frequency range. The resultant error in measurement of a real device's return loss is expressed as a +, − uncertainty and can be displayed as a ripple or repetitive waveform on top of the expected measured parameter. The better the source match the smaller the measurement uncertainty.

An uncorrected amplitude **level** variation, whether due to inherent generator performance or load sensitivity, over the frequency range of the device under test will introduce an equivalent value error.

Interrupted test sequences for changing frequency heads could introduce errors.

Sources with **harmonics** that are down less than 40 dB below the fundamental cause errors since detectors are voltage or power sensitive and can't differentiate between the various frequency components.

I.F. cram course



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The log IF strip is based on the Plessey SL1521, the simplest, easiest-to-use and least expensive wideband amplifier you can buy. It has a gain of 12 dB and an upper cut-off frequency of 300 MHz (our less expensive SL521 is available if you're working under 100 MHz). The SL1522 is two 1521's in parallel with a resistive divider for increasing the IF strip's dynamic range, while the SL1523 is two 1521's in series. With these devices, it takes just five cans and a single interstage filter to build a log IF strip at 160±15 MHz with a logging range of 90 dB, ±1 dB accuracy, −90 dBm tangential sensitivity and a video rise time of 20 nsec or less. It's reliable, inexpensive and field repairable, which is more than can be said about the hybrids or discretes you're buying now.

On the video output, our SL541 op amp matches the IF strip to your system, and lets you vary video sensitivity. It has the high slew rate (175 V/μsec), fast settling time (1% in 50 nsec) and high gain stability you need, with on-chip compensation so it's not tricky to use.

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By way of example[11] if the harmonics are only down 20 dB and the fundamental is at + 15 dBm the peak signal voltage will vary between .9 and 1.1 times the value corresponding to the value of the fundamental alone depending upon the relative phase of the two signals. Since the detector detects peak voltage at such a high power level its output will vary in the same way. Thus, the power display will show a variation between + 4.08 dBm to + 5.83 dBm, as the phase is varied. Below − 25 dBm the response is square law and the **power** simply adds. Accordingly for the example above but for operation at − 25 dBm the harmonic adds 1 percent to the true power so that the display would show − 24.96 dBm and not be phase-sensitive.

Unstable frequency sources can create tracking errors and additional unwanted modulation terms.

Signal Separation Device Errors

Signal separation or directional devices such as directional couplers, reflection bridges or hybrid networks have two characteristics that influence measurement accuracy: **Directivity** and **Test Port Match.**

Directivity Error

A directional device terminated in Z_0 is shown in Figure 7.

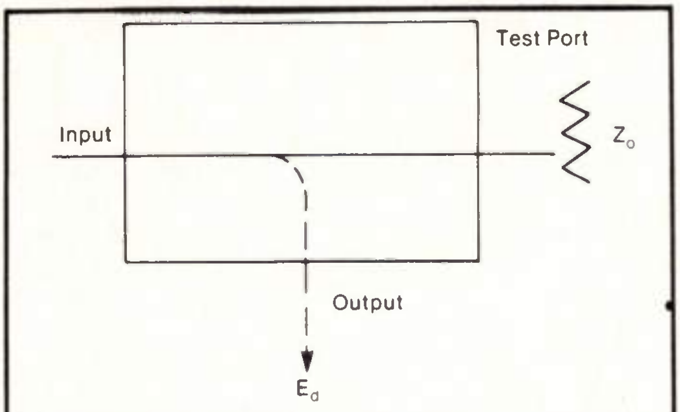

Figure 7. Directional Device terminated in Z_0.

All of the input is **not** absorbed in the perfect termination. A low level signal appears at the output. This (directivity) signal arises from deviations from ideal within the directional device and could have resulted from deviations from prescribed geometry, connector mismatches, or imperfect internal terminations. The summation of all these effects is called the directivity of the device and is expressed as dB below a full reflection as measured at the output port. Typical valves are 35 dB.

Directivity influences a measurement by adding an additional undesirable component of unknown phase relationship to the return loss component of the D.U.T. impedance. (This is similar to the source match error sequence of events previously discussed.) The following example illustrates the magnitude of the meaurement uncertainty this type of error produces. (Figure 8A).

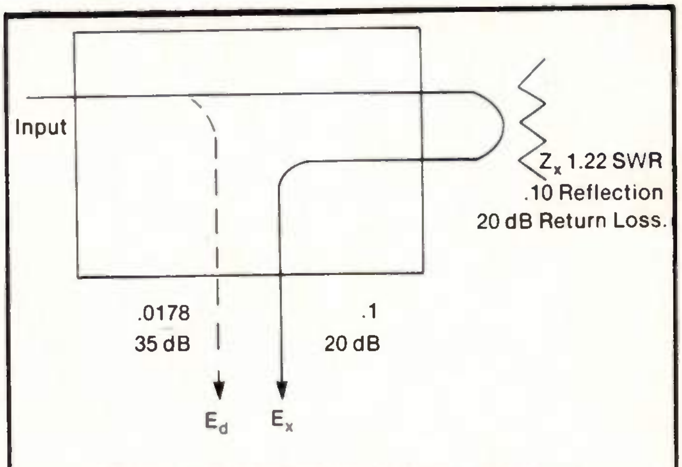

Figure 8A. Directivity Influenced Error.

The unknown being measured is a 1.22 SWR (20 dB return loss) which is only 15 dB larger than the signal due to directivity. The measured signal at the output port will vary from a minimum of − 18.58 dB to 21.70 dB down depending upon the phase relationship between Ex and Ed. This is the extent of the uncertainty — all caused by the finite directivity.

This uncertainty is greater for a lower SWR D.U.T. since the return loss and directivity components are closer in amplitude with unknown phase relationships, especially over a wide frequency range.

Test Port Match Error

If the test port, in Figure 8B, has other than 1.00 SWR a re-reflection will occur producing a signal component (E_2), of varying phase, that will add to the output (E_1) and create an error.

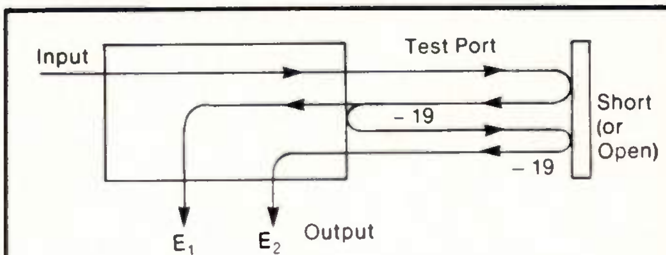

Figure 8B. Error due to Full Reflection.

As an example with a test port SWR of 1.25, 11.2 percent of the test signal will re-reflect. This reflected signal will be 19 dB below complete reflection. Arriving again at the short is fully reflected so that it arrives at the output differing by only 19 dB from the reference signal. Taking all phase relationships between E_1 and E_2 into account results in a − 0.92 to + 1.03 error variation.

However, in the real world, the load is other than a short and consequently absorbs part of the signal coming out of the test port. A little thinking reveals that this type of error rapidly converges, i.e. the error due to test port mismatch becomes smaller as the unknown becomes a better match.

These two errors are summarized in Figure 9.

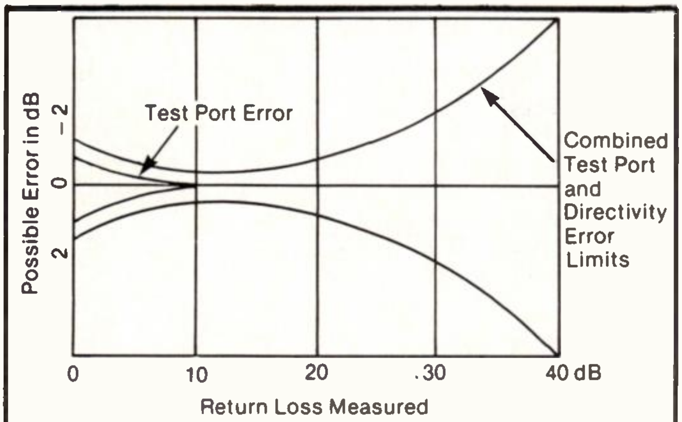

Figure 9. Error Limits of a Directional Device.

Notice how test port error converges for improving D.U.T SWR while the directivity error worsens.

Detector Errors

The Schottky diodes used in the detectors are non-linear devices approximately square-law at low signal levels to linear behavior at higher levels. If not compensated for, yet operated over a substantial power range, introduce significant errors.

Display and Signal Processing Errors

Scalar Network Analyzers that do **not** utilize modulated RF signals are sensitive to DC drift errors. DC level stability must be tight to prevent masking the DC component of low level (−50 dBm) detected signals.

Scalar Network Analzers that **do** utilize modulation schemes can introduce errors due to modulation asymmetry and additional insertion loss.

Adapter Errors

When the connectors on the D.U.T. differ from those on the reflectometer test port an adapter is often employed to make the measurement. However, the limitation to measurement accuracy caused by the adapter reflections is often severe. Suppose an adapter with 1.22 SWR (ρ = 0.10 and R.L. = 20 dB) is used in a reflection measurement. Since the adapter has a ρ of 0.10, it is not possible to measure D.U.T. ρ of less than 0.10. Otherwise, D.U.T. reflection may be obscured by the adapter reflection. Because the adapter reflection is a constant error quantity for all measured values ρ, it is added to the coupler directivity to obtain an effective reflectometer system directivity.

Corrective Measures

A corrective measures chart that considers the error producing situations already described is delineated in Table 1. (Page 32).

Expanded Dynamic Range

The dynamic range of a measurement system is limited at the high end by the sweeper output power or by the maximum level that can be detected accurately. The low end is limited by the detector sensitivity. Typical Scalar Network Analyzers have a 60 dB dynamic range; they can display signals as small as −50 dBm and as large as +10 dBm.

Two solutions for making measurements beyond the +10 dBm and −50 dBm range are to use an RF pre-amplifier to improve the detector sensitivity or to apply higher input power to the device under test when its attenuation drops the signal at the detector below −50 dBm. The expanded dynamic range technique uses the external leveling capability of the sweeper to prevent overpowering the detector when the insertion loss of the device under test is small.

Figure 10 shows the configuration for expanding the dynamic range using an amplifier after the device under test. The calibration power level is set such that the B detector sees +10 dBm, and the leveling loop attempts to hold this level constant. With 0 dB insertion loss, the level at the B detector is +10 dBm and the sweeper output power is −30 dBm. Moving to 40 dB insertion loss, the leveling loop holds the power at the B detector constant by increasing the source power to +10 dBm. The R detector therefore goes up 40 dB hence the ratio B/R shows 40 dB insertion loss. The main point is, with a 40 dB amplifier, 40 dB dynamic range is displayed by the R detector with the B detector held constant. An additional 60 dB dynamic range is displayed by the B detector with the R detector held constant, giving a total system dynamic range of 100 dB. If the amplifier is placed after the device under test, its critical characteristic is gain, and it needs to supply only +10 dBm. If the amplifier is placed before the device under test, it needs to

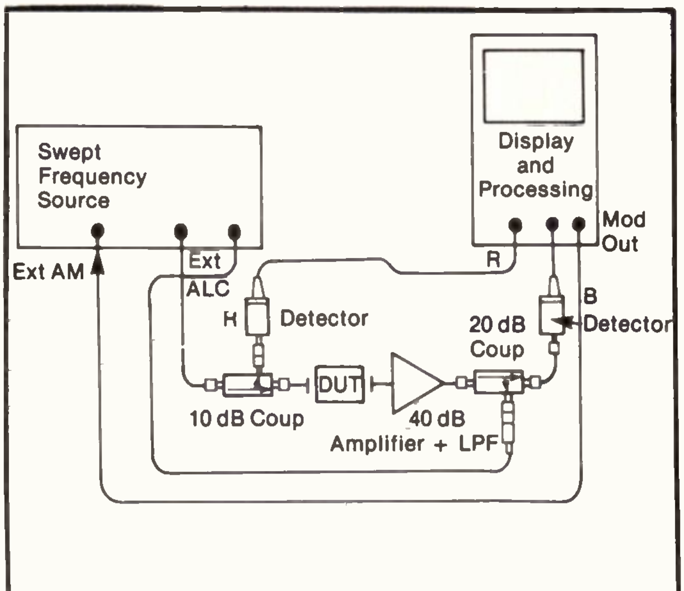

Figure 10. Increased Dynamic Range Using External Amplifier.

supply a power level equal to 10 dBm plus its gain.

Figure 11 shows the configuration for expanding the dynamic range using a high power source. The Network Analyzer is in a B/R ratio mode and the output power from the device under test is externally leveled using a 20 dB coupler and detector. During calibration the power level is set to provide +10 dBm to the B detector, and the leveling loop attempts to hold this level constant. When the attenuation of the device under test is 7 dB, the leveling detector calls for more power from the sweeper in its attempt to keep +10 dBm at the B detector. A +17 dBm source will be able to provide this power. Thus the power at the B detector stays at +10 dBm but the R detector sees a 7 dB increase. Since the Network Analyzer is displaying B/R, increasing the R level has the same effect on the display as decreasing B (i.e., in this example a 7 dB loss is displayed even through the power level at the B detector has not changed). For device under test attenuation of 10 dB the leveling signal requests +20 dBm from the source to maintain +10 dBm at the B detector. If the source cannot deliver more than +17 dBm then the leveling loop opens and the level at the B detector drops 3 dB. Therefore, the B detector provides its full 60 dB dynamic range while the +17 dBm source permits utilization of 7 dB of the R detector's dynamic range making possible a total dynamic range of 67 dB. To obtain substantial improvement in dynamic range using this method would require very high power sources. For example, 90 dB dynamic range would require a 10 W source.

Automation

What is automation with respect to Scalar Network Analyzers? The consensus of opinion is that there is **no** consensus. What one person considers automated another considers semi-automated. Groping

Table 1

Error Producing Situation	Corrective Measure	± Ramifications
(Sources)		
Mismatch	Series Pad Internal Leveling External Leveling Dual Coupler Isolator Use ratio technique.	Reduces dynamic range Minimizes number of connections. Reduces connector/cable mismatches. Frequency sensitive.
Level Variation	Internal Leveling External Leveling Use ratio technique.	(see above) (see above)
Interruptions	Wider coverage heads.	Greater Cost.
Harmonics	Use Tracking Filter Use Low Pass Filter Pad the reference detector.	Expensive Limits sweep bandwidth.
Frequency Instability	Phase-Lock to low frequency source.	Limited to CW mode only.
(Signal Separation Devices)		
Low Directivity	Use reflection bridge over coupler. Use magnified reflection technique.	Additional power loss. Requires special 4 port 50 ohm bridge.
Test Port Mismatch	Use both an open and short during calibration Use same connector as D.U.T. Use ratio technique.	Helps establish an accurate calibration reference. If before reference detector
(Detectors)		
Non-Linear Characteristic	Processor Compensation Circuits	Added complexity.
(Display and Processing)		
DC Drift	Built in thermocouple to offset temperature change. Low level adjust. Renormalize when temperature stabilizes.	
Modulation Asymmetry	Use ratio technique.	Eliminates Insertion/Return Loss error
(Adapters)		
Reflections	Use improved series. Careful placement in system. Minimize use.	If available. Digital Normalization might be effective then. Not always possible.

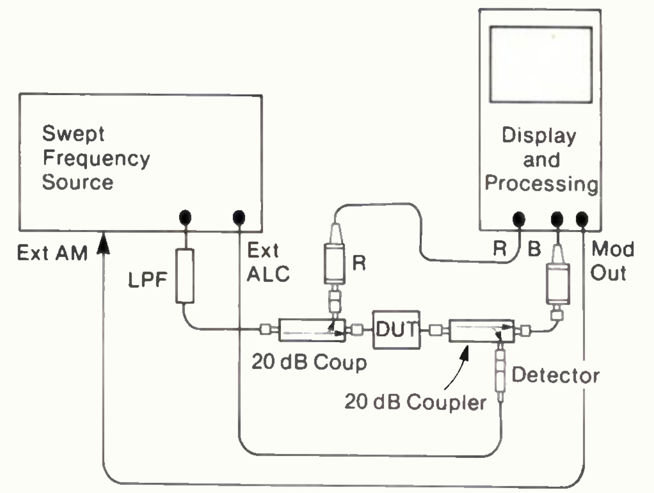

Figure 11. Increased Dynamic Range Using High Power Source.

for a definition it could be said that an automated piece of test equipment is one in which more than 90 percent of the front panel controls can be externally adjusted via a bus e.g. IEEE-488-1975, GPIB, HPIB, etc.

Why do we want an automated system? That perhaps is easier to answer. Such a system is not subject to operator fatigue — which gives us more consistent results in repetitive measurements. More devices and more parameters can be tested in less time. Results can be expressed in engineering units via on-line data processing. Greater accuracy is achievable since system errors can be measured automatically, stored and compensated for in the results. Data can be stored and transmitted to be utilized elsewhere whether in an adjacent piece of test equipment or half way around the world as in a global communications antenna performance monitoring system.

Prior to "automated" Scalar Network Analyzers an RF engineer would calibrate out his test equipment residuals, take a grease pencil mark its swept trace on a scope and be ready to begin testing the unknown device. Now, through the advent of **"Automemory" "short-cut keys" "storage-normalization"** or **"self-zeroing"** (as some of the main manufacturers call it) these, as well as other techniques is what separates an automatic from a manual system.

Automemory[12], for example, automatically removes errors in amplitude measurements so that the correct response of a tested device can be seen. It enables the storing of any response as a 0 dB reference against which future swept measurements may be compared. Automemory by means of automatic scale factor correction allows a reference response recorded with one sensitivity to be compared on any sensitivity.

"Short-cut keys," representative of another system[7], enables an operator, through push-button control of a controller's program to adaptively sequence, at his disgression, whatever steps he feels necessary to expeditiously complete a test routine of a D.U.T.

Storage-normalizers, as utilized in conjunction with a particular frequency response test set[6], is an outboarded unit that provides a completely flicker-free trace independent of the RF sweep rate. Its digital normalization capability allows subtraction of a digitally stored trace from the current measurement trace.

"Self-zeroing, self-calibrating" features of a microprocessor-controlled microwave multimeter forms the basis for another useful instrument in the RF designers arsenal.[4] The trend with this as well as the other described instruments is to take the "eccentricities" out of RF design measurements. They are intended to enable an operator with rudimentary knowledge of devices and instrumentation to follow a "prompted" sequence of procedures to arrive at consistent, repeatable test results.

What about IEEE-488-1975? As alluded to several times in this article mechanical, electrical and system interconnection between Scalar Network Analyzers and a computing or calculating controller for data manipulation purposes is achieved. The IEEE standard, Hewlett-Packard Interface Bus or General Purpose Interface Bus as it is commonly known consists of 16 active signal lines. It is grouped into three sets according to function: 8 data lines, 3 data byte transfer control lines and 5 general interface management lines. As mentioned previously there are **talkers** that transmits data, **listeners** that receive data and **controllers** that manage communications on the bus, primarily by designating which devices are to send or receive data during each measurement sequence. The controller may also interrupt and command specific actions within devices.

The beauty of compliance with the interface standard is that the RF design engineer can pick and choose the "finest" component instruments and be assured of compatibility among units made by different manufacturers.

Peripherals

Peripherals enhance and expand the capabilities of Scalar Network Analyzers. Several of them are described below:

Permanent records of all swept responses are possible with a **graphics plotter,** response recorder or X-Y plotter as it is variously known. Typical specifications include:

Plotting Area — 8" x 11"
Plotting Accuracy — ±0.016"
Repeatability — 0.016" from any given point.
Step Size — 0.0013" minimum
Speed — 10"/second in each axis
Storage-Normalizer is a versatile addition to several Network Analzers[10] that provides digital storage display and normalization that:
Refreshes a CRT at a flicker-free rate while updating stored data at the actual sweep rate.
Stores a system's residual response and automatically

Table 2

Warning: To fully appreciate and utilize these specifications — the OEM data should first be consulted. This is a guide only.

Model	Hewlett-Packard HP 8755	Narda 7000A	Pacific Measurements 1038/D14	Wiltron 5610
Frequency Range	10 MHz-18 GHz	10 MHz-18 GHz	1 MHz-18 GHz	10 MHz-18 GHz
Dynamic Range	60 dB	60 dB	60 dB	60 dB
Sensitivity	− 50 dBm	− 50 dBm	− 50 dBm	− 50 dBm
Channels	A, B	3 1/2 digit LCD	A, B	A, B
Resolution X Axis	40,000 points 10 MHz-18 GHz	52 points across band	512 points across band	10,000 points over 10 MHz-18 GHz
Resolution Y Axis	0.02 dB using 598313A	0.1 dB	0.1 dB	0.01 dB
Directivity	26-38 dB	40 dB	40 dB	40 dB
Generator Output Power	+ 10 dBm up to 18 GHz	Uses	Uses	+ 10 dBm up to 18 GHz
Harmonic Levels	25 dB down	outboarded	outboarded	30 dB down
Spurious Levels	40 dB < 2 GHz, 50 dB > 2 GHz	sweeper	sweeper	60 dB down (typical)
Accuracy*	1 dB	± 3% + M.U.	0.1 dB per 10 dB	1 dB
IEEE Bus Compatible	Requires Option	Requires Option	Requires Option 04	Yes

*The warning is especially true for this complex specification. Please refer to OEM data **carefully.**

displays the measurement data minus the reference (normalization).

Outputs its memory at a constant rate independent of the actual system sweep rate.

Thermal Printer is a 80 column printer that produces single-copy, page width, fully formatted, alphanumeric text, tables or simple plots at 240 lines per minute.

Desk Top Computer with over 23K byte of memory, utilized as a IEEE-488-1975 interface standard controller is available and described under "Automation".

Product Lines

Research revealed that eight U.S. manufacturers of test equipment produce (or produced) a system that could be considered a Scalar Network Analyzer. In Alphabetical order they are:

General Microwaves
General Radio (Gen Rad)
Hewlett-Packard
Narda
Pacific Measurements Inc.
Tektronix
Weinschel
Wiltron

Adding the following restrictions:

- Presently available
- Measures Insertion Loss, Return Loss and Power Level
- Displays Insertion Loss, Return Loss and Power Level Swept Responses
- Provides digitized stored memory capability
- Covers the .01 to 18 GHz frequency range
- Is IEEE bus compatible (either directly or with additional interface circuitry)
- Requires minimum number of interconnections

Under these restrictions Hewlett-Packard, Pacific Measurements Inc. and Wiltron have available units. (Narda's multimeter with an external display also meets these requirements.) They are:

Hewlett-Packard — Model HP8755 Frequency Response Test Set
Narda — Model 7000A Microwave Multimeter
Pacific Measurements — Model 1038/D13 Swept Frequency Measurement System
Wiltron — Model 5610 Automated Network Analyzer System

A chart "comparing" these units is detailed in Table 2. □

Footnotes

1. Alford and Weinschel in 1968.
2. Pacific Measurments Inc.
3. Digital Interface for Programmable Instrumentation.
4. Narda Microwave Multimeter Model 7000A.
5. P.M.I. Application Note 18.
6. HP8755 Frequency Response Test Set.
7. Wiltron Model 5610 Automated Scalar Network Analyzer System.
8. 100 kHz — 18 GHz: many manufacturers; 26.5 to 110 GHz Hughes Aircraft 44410H main frame and 41,000 series plug-ins (41011H, 41031H, 41051H and 41061H).
9. HP 11665B modulator.
10. HP 8750A storage normalizer used with HP 8507A, 8410, 8505A and 8755.
11. P.M.I. Application Note 12.
12. A registered trademark by Pacific Measurements, Inc.

Acknowledgements

I want to personally thank the following individuals for the patience and help they gave me in providing material, ideas and suggestions: Rob Coe of Narda; Bob Bathiany of Wiltron; Ralph Britton and Ed Mendel of P.M.I.; John Minck, Dean Abramson and Jim Simpson of H.P. and Ron Ramirez of Weinschel. I only hope I accurately perceived and utilized the information correctly.

Linear Amplifier Design: Some General Considerations

William Mueller
Communications Transistor Corp.
San Carlos, Calif.

Definition: *An ideal linear amplifier is one having a transfer function of the form $P_{out} = A\ Pin$ where A is constant over the entire useful power range (dynamic range) of the amplifier.*

This implies that the output waveform of such an amplifier is an exact, multiplied copy of the input waveform, i.e. that the amplifier is completely distortionless.

Unfortunately, in the real world there is no such thing as a completely distortionless amplifier. The exponential characteristic of a bipolar transistor's input diode causes distortion at low current levels. Nonlinearities in the transistor's current gain characteristics* (common emitter configuration) contribute significantly to third order products, especially at high current levels. Other transistor characteristics (e.g. depletion layer capacitance) also add nonlinearities.

The distortion caused by these nonlinearities can be classified into three categories:

- harmonic distortion
- intermodulation distortion
- crossmodulation distortion

When a single frequency signal or tone is applied to the input of an amplifier, components occur in the output signal not only at the fundamental frequency, but at integer multiples of it as well. These "extra" signals are called harmonic distortion products. When the input signal consists of more than one tone, the output waveforrn will also exhibit components at the sum and difference frequencies of the fundamentals. These are called intermodulation distortion (IMD) products. Crossmodulation distortion refers to the modulation of an "interference" signal being impressed on a "useful" signal operating at different frequency and is really a special case of intermodulation distortion.

The goal of the linear amplifier designer is to minimize distortion. Except in the special case of extremely broadband amplifiers (octave bandwidth or greater, where harmonic products fall within the pass band) harmonic products can be removed using low pass or band pass filtering techniques. IMD products that are too close to the desired frequency components to be removed by filtering, makes the designer's primary goal the minimization of IMD products.

Not limited to only bipolar transistors.

Applications

The two most common applications of large signal linear amplifiers are high power tansmitters using single sideband transmission and multi-tone data transmission systems used to convey information. In either case the desired signal is composed of simultaneous multiple tones. These give rise to intermodulation products, too close to the desired signal to be filtered, which distort the information content of the transmitted signal. For such systems it is essential to minimize this intermodulation distortion by using linear amplifier stages in the transmitter.

Measuring Linearity

There are many ways of specifying linearity. Perhaps the coarsest "figure of merit" in common usage is the compression point. It is a measurement, usually in dB, of the deviation of the output power waveform under large signal conditions from the small signal linear case defined by the transfer function $P_{out} = A\ P_{in}$. An output power is specified; then the compression (in dB) at this output power level is given by

$$compression = 10 \log_{10} \frac{A\ P_{in}}{P_{out}}$$

where P_{in} and P_{out} are measured in watts. The compression point provides a rough index of the maximum linear power available for AM type signals.

Many applications calling for linear amplifiers require a more precise measurement of linearity than is provided by specification of a compression point. In these cases a multiple tone input signal is used and nonlinearity is specified by measuring the resulting distortion products relative to some reference point.

In most cases this input signal takes the form of two tones of equal magnitude, very close to one another infrequency. The distortion product most commonly measured is the third order product d_3, also called IM_2 for the case of a two tone test. Two such products appear in the output, one at $2f_1 - f_2$, and one at $2f_2 - f_1$, where f_1 and f_2 are the frequencies of the fundamental tones. Whereas in theory these products are of equal magnitude, the gain slope of the amplifier being tested may cause slight variations in their amplitudes. If this is the case, an average value should be taken.

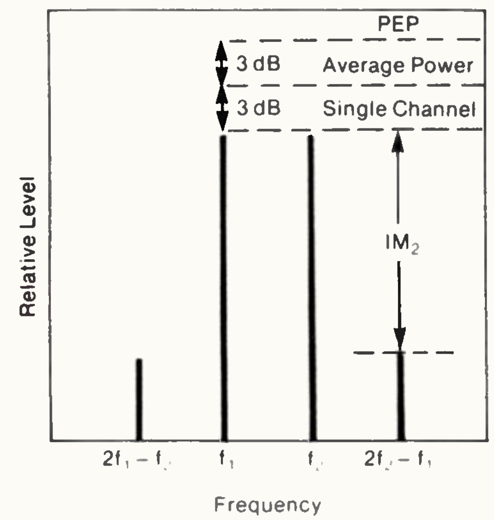

Figure 1. Intermodulation Factors with 2 Tone Method.

The magnitude of IM_2 is then specified in decibels below some reference point. Most commonly used is the power level of either fundamental tone, called the single channel level. The average power level, equal to the combined power of the fundamental tones, is another common reference level; it is also the power level that would be read using an average reading power meter. For two equal tones this level is 3 dB above single channel level. The third reference commonly encountered is the peak envelope power (PEP) of the fundamental waveform; for two equal tones this level is 6 dB above single channel level (See Figure 1). Care must be taken to explicitly specify which reference level is being used when measuring IM_2.

Occasionally instead of specifying the magnitude of the nth order distortion product d_n, an nth order intercept point will be given. Geometrically this is the point (P_{out} level) at which the linear projections of the curves P_{out} vs P_{in} intersect; algebraically it is given by the equation:

$$I^n = P_{out} + \frac{|dn|}{n-1}$$

where P_{in} and P_{out} are measured in dBm and dn is specified in dBc. For the third order intercept point this reduces to:

$$I^3 = P_{out} + \frac{|IM_2|}{2}$$

If IM_2 is referenced to the single channel level, then I^3 is typically 10 to 12 dB above the saturated output power of the amplifier being tested.

For certain applications the input waveform used for measuring distortion may become considerably more complicated. The television transmitter industry provides a good example of such a special case. Three unequal tones are used as the input signal; they are established as follows. A zero level (0 dB) is defined corresponding to the peak sync power level. An audio tone is set at −7 dB to model a 5:1 video to audio power ratio. A video tone is set at −8 dB, about halfway between black and white levels. A sideband carrier is set at −16 dB. Frequency separation is as shown in Figure 2. This signal is applied to the amplifier to be tested, and the level of the in band 3rd order distortion product (IM_3) is measured in dB below the peak sync level. The Swedish Post Office specification for television transmitters, perhaps the strictest European specification, requires that IM_3 be less than −54 dBc; typical design goals are for $IM_3 = -60$ dBc.

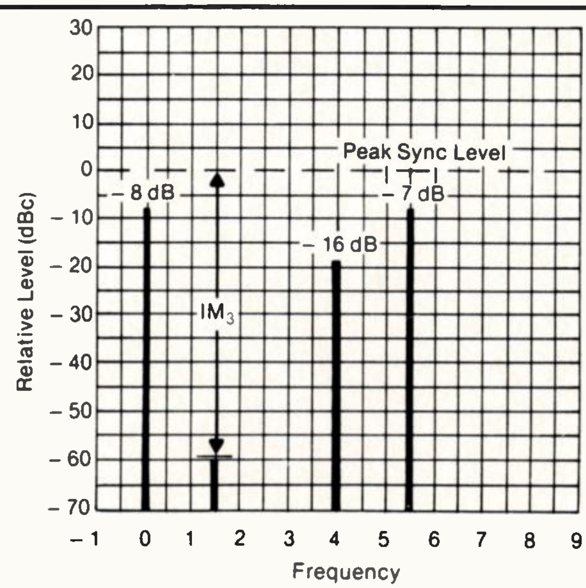

Figure 2. Intermodulation Factors with 3 Tone Method.

Whenever multiple tone signals are used to measure IMDs, special care must be taken to insure the "cleanliness" of the input signal. If IMD measurements are to be at all accurate, the distortion products of the input signal must be at least 10 dB below the level at which output measurements are to be taken. The drive source should be isolated from the amplifier under test by at least 10 dB. Adjustments in drive power level should be made using variable attenuators after the combining network to avoid frequency drifting of the input tones. Figures 3 and 4 show typical two and three tone test setups.

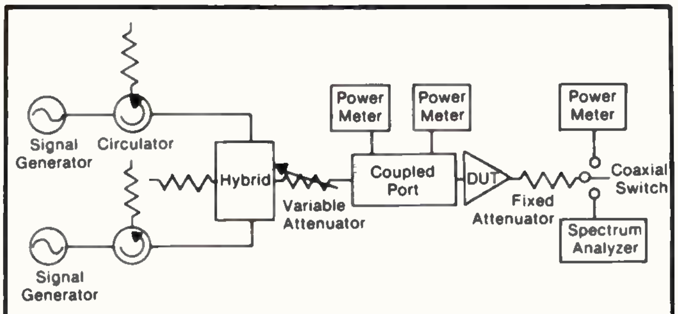

Figure 3. Linearity Test Setup, Two Tones.

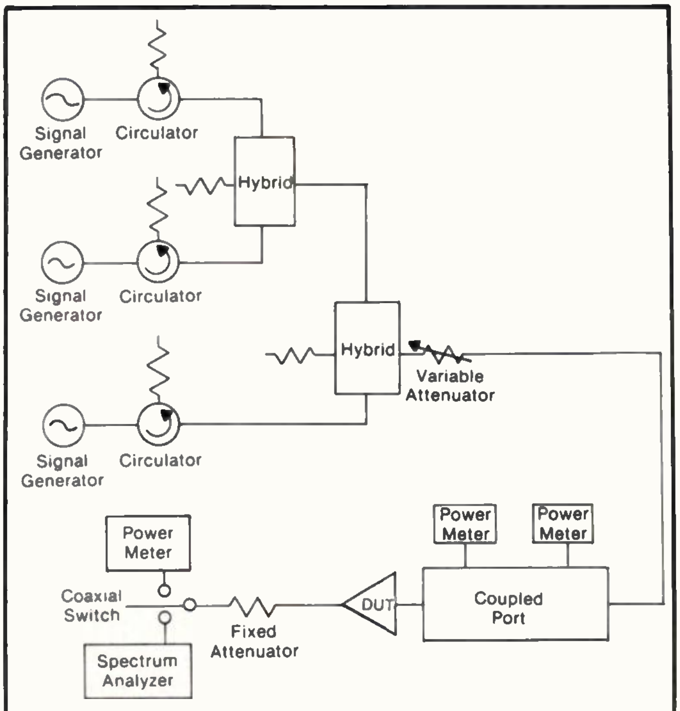

Figure 4. Linearity Test Setup, Three Tones.

The above measurements of linearity are not independent. As a rule of thumb, a 1 dB compression point indicates approximately the same output power level as does an IM_2 value (ref. single channel) of −20 dBc. Further, a relationship between IM_3 (ref. peak sync) and IM_2 (ref. single channel) can be derived theoretically and verified empirically. Assuming peak sync level (3 tone) equals PEP level (2 tone), then to a close approximation.

$$IM_3 = IM_2 - 14\,dB$$

For a complete derivation, see Reference 9. Remember that both IM_2 and IM_3 are negative numbers as defined above.

When two or more linear stages are cascaded, the worst case distortion level of the resulting multistage amplifier can be calculated if the gain and intercept point of each individual stage are known. The basic assumption made is that there is no in-phase coherency between the contributions of the individual stages. This yields quite accurate results for odd order intermodulation products, but is definitely in the worst case direction for even products due to both the distance in frequency from the fundamental that these appear and to the common use of push-pull techniques which result in phase inversion and cancellation of even products. For the general case:

$$i_T^n \exp\left(\frac{1-n}{2}\right) = \sum_{m=1}^{M} (i_m^n g_m) \exp\left(\frac{1-n}{2}\right)$$

Where i_T^n is the nth order intercept point of the total cascaded chain in mw.

i_m^n is the nth order intercept point of the mth stage of the chain in mw.

g_m in the power gain following the mth stage

$$\left(\frac{P_{oT}\,mw}{Pom\quad mw}\right)$$

M is the number of stages

n is the order of the intermodulation products under consideration.

Note that:

$$I_m^n = 10 \log_{10} \frac{i_m^n}{1mw}$$

Where:

I_m^n is the nth order intercept point of the mth stage expressed in dBm.

For the special case of the third order intercept point of two cascaded stages, the above equation reduces to:

$$\frac{1}{i_T^3} = \frac{1}{i_I^3 g_I} + \frac{1}{i_2^3}$$

converted to dBm this becomes:

$$I_T^3 = I_2^3 - 10 \log_{10} \left[1 + \frac{1}{g_1} \frac{i_2^3}{i_I^3}\right]$$

These equations indicate that for a two stage amplifier with second stage gain of 6 dB and equal distortion in both stages, the power level achievable by the entire amplifier will be 1 dB less than what the PA (Power Amplifier) stage alone could achieve at the same distortion level. If the contribution of the first stage is to be kept down to around 1%, its IMD's would have to be at least 14 dB lower than those of the PA stages. For this reason many designers would use the same sort of transistor for both driver and PA stages — the driver would then theoretically have IMD's 18 dB lower (for a 6 dB gain transistor) than the PA stage.

Circuit Considerations

The nonlinearities of the voltage and current saturation regions and the turn-on region of large signal power transistors are primary sources of distortion. Consideration should be given to these regions when designing any linear amplifier, and special care should be taken to avoid them when selecting load impedances and bias points.

The circuit designer has three primary bias conditions to choose from. Class C operation is characterized by a collector conduction angle of less than 180°, and is achieved by reverse biasing the base-emitter junction of the transistor. Its primary characteristics are as follows:

- poor linearity (IM_2 ref. single channel at −10 to −20 dBc typical)
- poor dynamic range (5 to 10 dB at 1 dB compression points)
- lowest gain
- highest efficiency (50% to 70% typical broadband; up to 85% narrow band)
- highest output power from device

The high efficiency, zero static drain and consequent reduced thermal problems of this mode of operation make it very popular for many RF applications; the high distortion and small dynamic range make it impractical for most linear designs, however.

Class AB operation refers to a collector conduction angle somewhat greater than 180°, and is achieved by slightly forward biasing the device's base-emitter junction. This forward biasing eliminates the cross-over distortion inherent in the transistor's turn on region, and greatly improves linear performance. Some typical characteristics of class AB operation are:

- good linearity (IM_2 ref. single channel at −20 to −40 dBc (typical)
- good dynamic range (30 dB typical)
- improved gain (approximately 1.5 dB above class C operation)
- medium efficiency (30% to 50% typical for broadband operation)
- output power only slightly below class C capability.

This mode of operation is a good compromise between high efficiency and linearity, and is used in nearly all linear applications not having particularly stringent IMD requirements.

In class A operation, the base-emitter junction is fully forward biased at all times, i.e. the collector conduction angle equals 360°. An amplifier biased in this manner will display the following characteristics:

- best linearity (IM_2 ref. single channel −30 to −50 dBc or better)
- best dynamic range (typically 50-70 dB)
- best gain (approximately .5 dB above class AB operation)
- poor efficiency (10% to 20% typical)
- output power capability approximately 6 dB below class C level.

The thermal problems accompanying the inefficiencies of this class of operation make it useful only where strict linearity requirements are encountered (eg. data transmission systems).

To achieve a class AB bias, the base-emitter junction of the device under test must be slightly forward biased so as to draw enough static current to avoid operation in the turn-on region. Typically, 50 to 100 mA will suffice. The optimum value of the effective DC bias source impedance can then be calculated from the equation:

$$R_S = \frac{h_{FE}(V_{BE_1} - V_{BE2})}{I_{C2}}$$

Where h_{FE} is the DC β of the transistor

V_{BE_1} is the base-emitter voltage required to produce the desired static current

V_{BE2} is the base emitter voltage which produces minimal distortion at the maximum desired power output level

I_{C2} is the DC collector current drawn at the maximum desired output power level

Typical values for R_S are on the order of one ohm.

Once the bias point has been established, thermal stability must be achieved. The DC base to emitter voltage of a transistor decreases as the temperature of the junction increases. Therefore, if the bias circuit holds the base voltage fixed in a device operating in the common emitter configuration, as the transistor heats up during normal operation more and more collector current will be drawn. This will further heat the transistor, in turn drawing still more collector current with the process continuing until the device is driven into thermal runaway. The bias circuitry must, therefore, provide some means of temperature compensation to avoid the destruction caused by such thermal instability.

References describe several class AB bias circuit techniques; most are either very complicated or have problems maintaining DC thermal stability when used with RF transistors having high values of h_{fe}. To surmount these difficulties several companies developed a semiconductor device which when used properly, provides excellent temperature tracking for DC stability in a significantly simplified bias circuit. Further, no supplemental emitter resistance is needed to ensure DC stability.

For class A operation a bias circuit which achieves thermal stability through feedback techniques is recommended. One such circuit is shown in Figure 5. In this circuit the PNP device acts as a current source for the base of the RF transistor. R_1 controls the minimum collector current that can be drawn by the RF device. R_2 establishes the bias point at which the PNP device is operated. R_3 serves primarily as a power dissipater, lowering the thermal load placed on the PNP device. R_4 reduces the effects of leakage currents on the bias of the RF device; R_5 sets the collector current drawn by the RF device. The diode is included to thermally track the junction of the PNP device.

The circuit functions as follows. As the RF device heats up during normal operation, it begins to draw more collector current. This increases the voltage drop across R_5, and consequently shuts down the emitter voltage on the PNP transistor. This in turn reduces the collector current of the PNP transistor. As this current supplies the base of the RF device, the collector current of the RF device is also lowered. Thus a stable bias point is achieved through feedback.

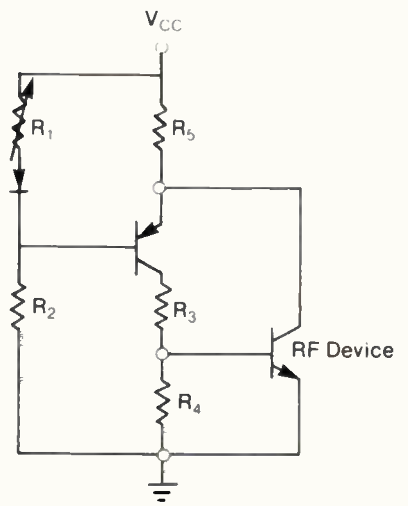

Figure 5. Class A Bias Circuit Using Feedback (RF blocking and matching circuitry omitted.)

When selecting a bias point for class A operation, Ic is either selected empirically or chosen from data supplied by the manufacturer. A typical class A bias point would be Ic = 0.4 Ic max, Vce = 0.8 Vce max, Ic max and Vce max representing the maximum safe DC operating limits of the transistor.

As mentioned above, correct choice of load impedance will keep the signal swing of the transistor clear of the saturation regions, and so help minimize distortion. For a device in class C or class AB operation, the collector voltage swing is very nearly given by:

$$V_C \, or \, AB \cong \frac{V_{CC} - V_{sat}}{\sqrt{2}}$$

A device in class A operation has a much more limited collector swing; a closer approximation for this case would be:

$$V_A \cong \frac{V_{CC} - V_{sat}}{\sqrt{7}}$$

From the definition of power:

$$P_{out} = \frac{V2}{R_L}$$

Rearranging terms and incorporating a correction factor K to correct for efficiency constraints (K typical = 0.8) one has

$$R_L = K\frac{V2}{P_{out}}$$

Therefore, for class AB operation:

$$R_L \cong K\frac{(V_{CC} - V_{sat})^2}{2\,P_{out}}$$

and for class A operation:

$$R_L \cong K\frac{(V_{CC} - V_{sat})^2}{7\,P_{out}}$$

To complete the model for load impedance, the effective output capacitance of the transistor (Cob) and the package lead inductance (Lc) should also be considered, as in Figure 6. Cob can be directly

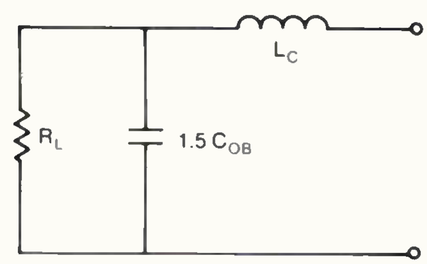

Figure 6. Effective Load Impedance.

measured using a capacitance meter and a bias supply; often a typical value will be included on the manufacturers spec sheet. Lc depends on package dimensions and the details of bonding, but is typically in the range .3 nh to 1nh.

The input impedance should either be transformed up to the characteristic impedance desired for the system cable connections, or to the desired impedance used for loading the previous stage, depending on system requirements.

Negative feedback, either collector to base or emitter degeneration, can also be used to trade amplifier efficiency, gain and stability for improved linearity.

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Index of Articles

A wealth of practical and theoretical information has appeared in the ten issues of *r.f. design* since the first issue in November 1978. An index to these articles is included. The categories are **Amplifiers, Antennas and Propagation, Components-Active, Components-Passive, Measurements and Test Equipment, Systems, and Techniques.** Future indices will be included on a more regular basis.

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Calculator Design of Microstrip Traps	Design of microstrip filters using an HP-19C/29C calculator program.	A Przedpelski	7907	21
Microstrip Translator Program for your HP-67	The translation of an existing microlayout to a different substrate material using an HP-67 calculator program.	F Hauer	7911	32
7 Element, 50 ohm Chebyschev Filters using Standard Value Capacitors	By proper selection of reflection coefficient and cutoff frequency standard value capacitors can be used in filter design.	E Wetherhold	8002	26
Design with Pin Diodes, Part I	Pin diode theory, applications and performance data are described.	G Hiller	7903	34
Design with Pin Diodes, Part II	Pin diode attenuator and phase shifter designs and incidental distortion from them are detailed.	G Hiller	7905	40
Design of VHF Quadrature Hybrids, Part I	Simple design equations for a narrow band quadrature hybrid are derived using scattering parameters.	CY Ho	7907	49
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Directional Couplers	Theory and design charts for directional couplers are given. A five-way power splitter is detailed.	A Paolantonio	7909	40

Measurements and Test Equipment

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Transceiver Testing	A single instrument (HP8901A Modulation Analyzer) provides transceiver performance analysis and measurement.	JL Minck	7811	24
Frequency Counter Cornucopia	A survey of frequency counter specifications.	RF Staff	7901	49
Spectrum Analyzer Analysis	The manufacturers discuss the capabilities and specifications needed to accomplish various frequency measurements.	M Engelson	7905	14
Spectrum Analyzer Performance	The meanings and interrelationships of spectrum analyzer specifications are detailed.	R Fowler	7905	28
Sampling Uses for RF Measurements In Instrumentation	Examples of Extended time sampling, sample and hold circuits and their relationship to instrumentation is given.	D Gittens	7905	48
Advance Instrument does RF Impedance Analysis	HP4191A measures 14 impedance terms: $\lvert Z\rvert$, $\lvert Y\rvert$, $\lvert\Gamma\rvert$, θ, R, X, G, β, Γ_x, Γ_y, L, C, D and Q from 1-1000 MHz.	J Mathews	7911	14
Measuring Generator Output Impedance	The measurement of the complex output impedance and reflection coefficient with a line stretcher and RF millivoltmeter.	R Lafferty	7911	27
Signal Generator Specifications, Part I	Review of important generator specifications.	R Oeflein	8001	14
Signal Generator Specifications, Part II	Output level, Modulation, Speed and Control relationships to signal generator performance are discussed.	R Oeflein	8002	11
RF Signal Generator Specification Matrix	A chart of 8 manufacturers signal generators models versus performance parameters is given.	RF Staff	8001	24
Scalar Network Analyzers	An historical, theoretical and operational perspective is developed for this broadband measuring instrument.	RF Staff	8003	21

Systems

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An AM-FM Digital Tuning System	Design utilizing a Digitally Tuned Radio Controller, Phase Lock Loop Synthesizer and Driven Display ICs is described.	TB Mills	7811	31
25,000 Filters in One	Real time analysis over large IF bandwidths using dispersive delay lines in a compressive IF receiver.	W Crofut	7909	20
Phase Lock Loops	Includes the effects of finite gain, finite bandwidth op-amp in a 5th order loop transfer function.	A Przedpelski	7909	24

Techniques

Title	Description	Author	YRMO	Page
Coax provides DC Power for Receiver	A means of providing DC power to outboarded circuits while simultaneously receiving RF energy on the same coax.	RF Staff	7811	50

EMI Filter Catalog

Stanford Applied Engineering, has available a new 14 page electromagnetic interference (EMI) filter catalog. Included are complete details on the electrical and performance specifications on the company's line of general applications, high performance, switching transient, three phase, connector, medical and extended range filters. Enlarged drawings of electrical schematics, performance graphs, and case styles are clearly defined along with easy to read ordering information. An in depth discussion on the applications and selections of EMI filters is included for customer convenience. All SAE filters in the catalog are designed to meet UL and CSA requirements.

Contact C. Jean Littrell, Stanford Applied Engineering, 340 Martin Ave., Santa Clara, CA. INFO/CARD #100.

Biological Effects and Medical Applications of Microwave Radiation

A collection of state-of-the-art research papers on the biological effects and medical applications of electromagnetic (EM) energy in the radio-frequency (RF) spectrum appears in the January 1980 *Proceedings of the IEEE.* The *Proceedings* is a monthly publication of The Institute of Electrical and Electronics Engineers, the world's largest professional engineering society, and contains papers of broad significance in all areas of electrical science and technology.

Four papers describe the extent of general population exposure to radio-frequency EM fields from broadcast signal fields and telecommunications systems and the extent of occupational exposure in various industrial processes.

Among other specific topics covered in the issue is a review of the epidemiologic studies of biological effects of microwaves by Dr. Charlotte Silverman of the Bureau of Radiological Health of the US Food and Drug Administration. This paper includes descriptions of a study of US naval personnel occupationally exposed to radar, and a study of American Embassy personnel in Moscow exposed to long-term microwave irradiation by the Russians. Neither of the two studies demonstrated differential health risks associated with presumed exposure to microwave radiation. Dr. Silverman also discusses current knowledge of the effects of microwave radiation on vision, behavior and the nervous system and suggests avenues for further research.

In another paper Dr. Donald I McRee of the National Institute of Environmental Health Sciences presents an overview entitled "Soviet and Eastern European Research on Biological Effects of Microwave Radiation." Because Soviet and Eastern European studies tend to omit details concerning the experimental design and exposure techniques, many of their conclusions have been ignored in the US. However, this literature does indicate a large number of bioeffects at low exposure levels (below 10 mW/cm^2), which has generated concern. Dr. McRee's paper summarizes research results since 1976, in which increased data using long-term microwave exposures at power density levels below 10mW/cm^2 have been reported. Replication by a US research group of Soviet results at extremely low levels (500$\mu W/cm^2$) emphasizes the need for performing long-term, low-level microwave bioeffects research by Western investigators and the necessity of evaluating seriously the results of Soviet and Eastern European research.

Contact IEEE Single Sales Department, 445 Hoes Lane, Piscataway, NJ 08854; the price is $5.00 for members and $10.00 for nonmembers. INFO/CARD #101.

RF Communications

Practical RF Communications Data for Engineers and Technicians by M.F. "Doug" DeMaw is a practical no-nonsense compilation of useful charts, circuits and explanations that has a cohesiveness of content and simplicity of language that only a person with a deep understanding of electrical theory and years of experience could write. Doug DeMaw, design engineer and senior technical editor for the American Radio Relay League has compiled a 256 page text of useful data divided into the following categories: Ferromagnetic Devices, Networks for Semiconductor Circuits, Transmitter Design Basics, Receiver Design Basics and Practical Communications Antennas.

It is published by Howard W Sams & Co. and sells for $8.95. INFO/CARD #102.

All about Transmission Lines

Electrical Characteristics of Transmission Lines by Wolfgang Hilberg represents a uniform, accurate and systematic approach to computing the characteristic impedance of transmission lines. Using a system based on stereographic projection and combining that with the known methods of the corresponding diagrams, Hilberg shows how the reflections and utilization of properties of complementary lines, as well as most of the characteristic impedance formulas contained in the known tables, can be systematically obtained simply and intuitively from the formula of a single line.

The text is divided into 4 parts: A) The general calculation method and the determination of the characteristics of a series of transmission lines; B) Tables of characteristic impedance; C) Diagrams; D) References concerning transmission line calculations.

The second part by itself is an extensive and useful reference volume. It is published by Artech House Books and sells for $31.00. INFO/CARD #103.

Solid State Microwave RF And Microwave Power Transistor Catalog

Solid State Microwave, a division of Thomson-CSF, has just released its new catalog offering a complete line of RF and Microwave power transistors for communications, radar, and avionics applications.

These devices are characterized from 2.0 MHz to 2.0 GHz for most FM, AM, pulse, sideband and linear applications. Also many gold metallized broadband internally matched structures are offered.

Contact: Jeff Holmquest, Solid State Microwave, Montgomeryville Industrial Center, Montgomeryville, PA 18936, (215) 362-8500. INFO/CARD #104. □

When quality counts

Do not be fooled by the low prices, these brand new lab quality frequency counters have important advantages over instruments costing much more. The models 7010 and 8010 are not old counters repackaged but 100% new designs using the latest LSI state-of-the-art circuitry. With only 4 IC's, our new 7010 offers a host of features including 10 Hz to 600 MHz operation, 9 digit display, 3 gate times and more. This outperforms units using 10-15 IC's at several times the size and power consumption. The older designs using many more parts increase the possiblity of failure and complexity of troubleshooting. Look closely at our impressive specifications and note you can buy these lab quality counters for similar or less money than hobby quality units with TV xtal time bases and plastic cases!

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7010 *7010.1	145.00 225.00	600 MHz	9	5-20 mV	10-30 mV	20-40 mV to 600 MHz	1-10 mV	(3) .1, 1, 10 SEC	.1Hz	1 Hz	10 Hz 600 MHz	1 PPM 0.1 PPM	10 MHz	YES OPTION $25.	YES OPTION $15.
8010 *8010.1	325.00 405.00	1 GHz	9	1-10 mV	5-20 mV	10-25 mV	1-10 mV	(8) .01-20 SEC	.1 Hz	1 Hz	10 Hz 1 GHz	1 PPM 0.1 PPM	10 MHz	YES STD	YES OPTION $39.

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#Ni-Cad-701	Ni-Cad Battery Pack & charging circuitry Installs inside unit	$ 15.00
#EC-70	External Clock input, 10 MHz	$ 25.00
#CC-70	Carry Case, Padded Black Vinyl	$ 8.95

MODEL 8010

#8010	1 GHz Counter - 1 PPM TCXO	$325.00
#8010.1	1 GHz Counter - 0.1 PPM TCXO	$405.00
#8010.1-13	1.3 GHz Counter - 0.1 PPM TCXO	$495.00

OPTIONS

#Ni-Cad-801	Ni-Cad Battery Pack & charging circuitry Installs inside unit	$ 39.00
#CC-80	Carry Case, Padded Black Vinyl	$ 9.95

ACCESSORIES

#TA-100	Telescope Ant with Right Angle BNC	$ 9.95
#P-100	Probe, 50 ohm, 1x	$13.95
#P-101	Probe, Lo-Pass, Audio Usage	$16.95
#P-102	Probe, Hi-Z, General Purpose	$16.95

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[illegible]

Siliconix Introduces the First Push-Pull RF VMOS Power FETs

Recognizing that many RF designs are push-pull, Siliconix is expanding its line by introducing three push-pull RF power VMOS devices for operation in the 2 to 200 MHz range. The Siliconix DV1112 delivers 120 watts minimum and the DV1111, 80 watts minimum. Also available is the DV1110, a 40 watt driver.

An application for this new series of push-pull configured VMOS power amplifier devices is high-gain, low-input VSWR amplifiers in the 30-88 MHz military communications band. All of the new power VMOS devices deliver rated output at 28 volts and provide a minimum power gain of 10 dB at 175 MHz. They can each be operated in Class A, B, or C and are well suited for a variety of broadband RF amplifier applications.

The well known advantages of the push-pull configuration, such as even-order harmonic suppression, are now added to the recently-introduced features of VMOS such as high power at high efficiency, low noise figure, absence of thermal runaway, ability to withstand infinite VSWR, as well as the ability to operate with simple bias circuitry. VMOS also is free of current hogging. Additionally, saturated output power is constant with frequency — unlike bipolars which have decreasing saturated output power with increasing frequency.

With VMOS baseband noise is 10 to 15 dB lower than in comparable bipolar devices and lends itself readily to broadband designs because of the relatively low inherent feedback capacitance.

An advantage to the push-pull package is the reduced amplifier size and the ability to directly connect impedance matching transformers to the devices.

These new push-pull transistors are easily broadbanded within the VHF bands. The lower frequency limit is governed only by the transformer design.

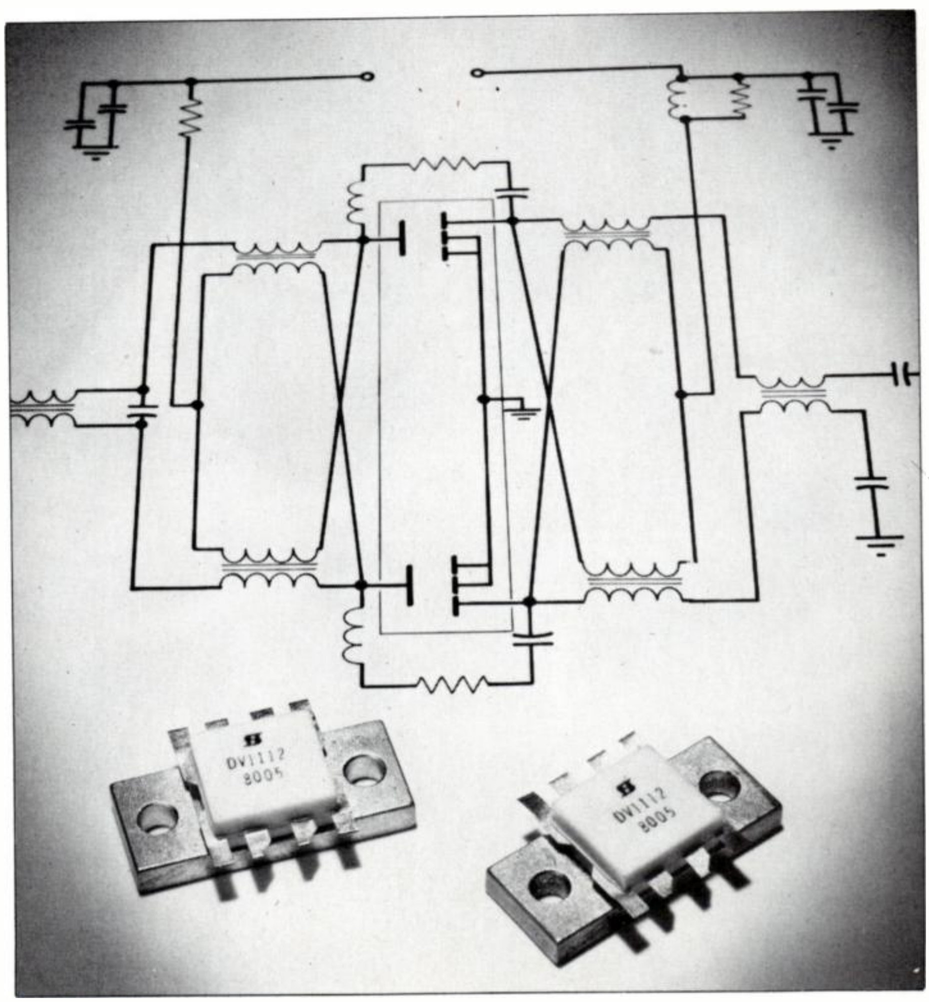
The following schematic represents an amplifier which produces 100 watts of power output ±0.2 dB from 30 to 90 MHz. Amplifier gain is 10 dB and efficiency is 64 percent across the band.

,Because the transconductance is virtually constant over wide frequency excursions, the input loading requirement also remains relatively constant within the hf region. Thus the input VSWR is quite stable over a broad bandwidth.

The inherently small feedback within the VMOS device, translates into the input impedance remaining virtually independent of reflective load impedance so that it is governed by the input matching circuit. Virtually no external feedback is required to ensure stability. Thus overall efficiency is high and out-of-band stability is significantly enhanced. Low external feedback also means that gain is flat across a very wide bandwidth.

These RF VMOS devices exhibit a reverse gain of at least −35 dB such that variations in output loading have little effect upon the input of the amplifier. VMOS power amplifier circuits exhibit the ability to withstand 20:1 VSWR at any phase angle.

With constant Rds(on). VMOS designs develop a leveled saturated output over a broad range of frequencies.



100 watt, 30-90 MHz amplifier.

New from THOMSON-CSF...

65 Watts Broadband 440–512 MHz

RF power transistor for UHF land mobile applications

- **All Gold Construction**
- **∞:1 VSWR at High Line**
- **Internally Matched**
- **Easily Broadbanded**

Thomson's new SD1499 internally matched 65 Watt 12.5 Volt RF power transistor was designed specificially for your next 100 Watt UHF land mobile radio design. This device produces a minimum of 65 Watts of output power across the full 440 to 512 MHz band. The SD1499 incorporates diffused emitter resistor ballasting and will withstand infinite VSWR loading at high line collector voltage (15.5V) conditions.

The SD1499 is a high gain device supplied in a low inductance .400 square package utilizing gold metallization and gold bonding wires.

For more information or samples, call Bob Tyson at (215) 362-8500. . .and get your next UHF design up to power.

Typical Parameters (V_{cc} = 12.5 V_{dc})

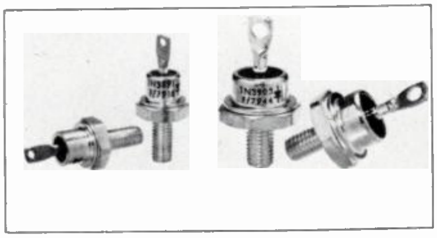
P_{out}	P_{in}	Z_{in}	Z_{cL}	η_c
70 W	22 W	1.25 + j .75 Ω	1 – j 1.5 Ω	60%

THOMSON-CSF

THOMSON-CSF COMPONENTS CORPORATION
SOLID STATE MICROWAVE DIVISION
MONTGOMERYVILLE INDUSTRIAL CENTER/MONTGOMERYVILLE, PA 18936 (215) 362-8500/TWX: 510 661 6548

Fast Recovery Rectifier Diodes

The Ferranti family of fast recovery rectifier diodes have voltage ratings of up to 400 volts, an average current



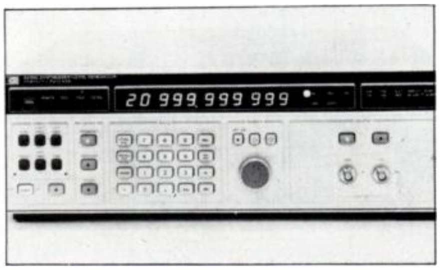
rating of 6 amps to 30 amps, and a typical reverse recovery time of 100 nanoseconds.

Contact Ferranti Electric, Inc., Commack,, N.Y. 11725. INFO/CARD #127.

Microprocessor-Based Frequency Synthesizer

This new 10 Hz to 20.9 MHz frequency synthesizer using microprocessor control is designed to be used as a precision signal source for telecommunications testing as well as traditional synthesizer applications. With a resolution of 1 μHz for frequencies less than 100 kHz and 1 mHz for frequencies above 100 kHz and amplitude accuracy of ± 0.05 dB with ± 0.1 dB flatness, this Hewlett-Packard model 3336C Synthesizer/Level Generator is useful for general purpose applications as well as frequency division multiplex (FDM) testing.

The 3336C can be used to test filters, mixers, phase detectors, attenuators, modulators or wherever a



high resolution, precision test signal is needed. Other communications applications include testing low-density carrier, radio baseband, high density cable carrier and R&D and production tests. For automated instrumentation systems, the model 3336C includes the HP-IB (IEEE-488) interface bus as a standard feature.

Frequency, amplitude and phase can be selected or incremented manually or automatically as the operator wishes. Low integrated phase noise (-60 dB over a 30 kHz bandwidth) is obtained from a break-through — HP's Fractional-N technology. For example, -70 dB phase noise in a 3 kHz band, 2 kHz from the carrier reduces noise interference from adjacent channels significantly.

Contact Inquiries Manager, Hewlett-Packard Company, 1507 Page Mill Road, Palo Alto, California 94304.

Noise Generator

A new noise generator, the NOD-102-1A has been added to the Micronetics noise product line. Covering a range of 500 MHz, the NOD 102-1A offers the following specifications bias +24V ± 0.5 percent; current less than 125 mA; frequency .005 to 500 MHz ± 1.5 dB (max.); output 0 dBm; Z load 50 ohm; terminations: noise — SMA (F), bias-solder lug; case 2" x 6" x 1" (exclusive of terminations), solder sealed steel. T operating 0° to 60° C.

Contact Micronetics, Inc., 36 Oak Street, Norwood, N.J. 07648. Circle INFO/CARD #125.

Low AM Radiation Power Amplifier

A new stereo power amplifier, featuring low AM radiation for use with compact radios, has been developed by National Semiconductor Corporation.

Known as the LM1896, the new device has a unique compensation scheme which allows increased sensitivity and signal-to-noise performance when used in AM radios.

The LM1896 reduces this radiation problem, and can increase the usable sensitivity over conventional designs by as much as 9 dB. In addition to low radiation and noise, the LM1896 has noise-free turn-on, and features low supply operation to 3V. Distortion is typically 0.1 percent at 1 kHz for output power of 5 watts.

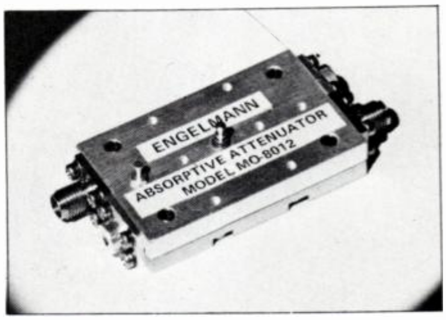
Contact National Semiconductor, 2900 Semiconductor Drive, Santa Clara, California 95051. INFO/CARD #123.

8-12 GHz Absorptive Modulator

Engelmann Microwave is now offering an 8-12 GHz Absorptive Modulator with four types of input tuning options. They include: TTL input "on-off" modulation; Analog current input; Linearization and Digital stepped (1 dB steps TTL).

The basic RF specifications of the model MO 8012 include a typical dynamic range of 70 dB and a maximum insertion loss of 3.0 dB, with a maximum VSWR of 1.7 under all conditions.

Switching speeds of the Absorptive Modulator are from 30-50 Nanoseconds

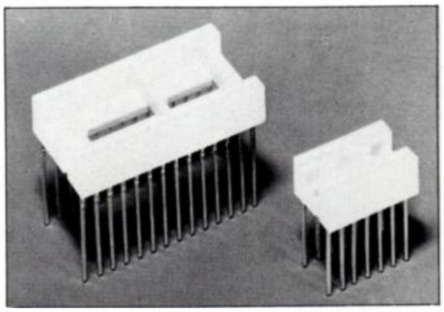


with a 50 watt peak and 1 WCW power handling capacity. Size of the model MO 8012 is approximately 3" x 2.5" x 0.75". The operating temperature range is -55 to $+95$°C with accuracy versus temperature rated at 0.03 dB/°C.

Contact Engelmann Microwave Company, Skyline Drive, Montville, N.J. 07045. INFO/CARD #117.

Wrap-Pin Socket

Robinson-Nugent, announced an additional low cost socket series. This new ICY series of wrap/pin DIP sockets, priced 25 percent lower than the standard "professional quality" R-N sockets, utilizes selective gold plating and a



different contact material (CA-275) to achieve the drastic price reduction. Available in 8 thru 40 pin sizes.

Contact Robinson-Nugent, 800 East Eighth Street, New Albany, Indiana 47150. INFO/CARD #121.

Rechargeable Sealed Lead-Acid Batteries

A new family of rechargeable sealed lead-acid batteries is now available from the Electronic Components Division of Panasonic Company. Designated as "LCR Series," the family consists of five 6-volt batteries.

Compared with conventional lead-acid battery systems, the LCR Series units offer greatly improved overdischarge cycle life: If the conventional lead-acid battery will withstand 20 extremely deep discharge cycles, the LCR Series battery will withstand over 100 cycles. Besides their ability to withstand overdischarge, they can also

withstand considerable overcharging, up to 1.2 Ah for a 3-Ah battery (compared to 0.2 Ah for a conventional unit). In addition, batteries that are put away discharged, can be retrieved after a period of storage, in most cases, simply by recharging.

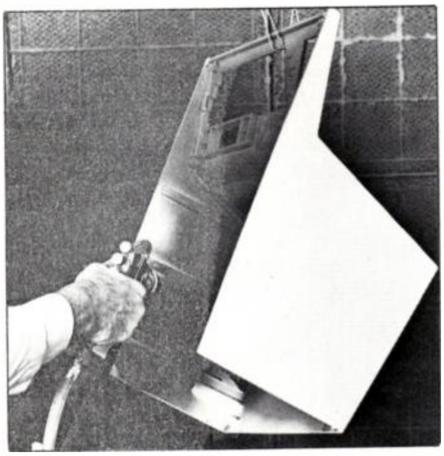
Contact Panasonic, One Panasonic Way, Secaucus, New Jersey 07094. INFO/CARD #119.

Sprayable Conductive Coating

Xecote, a dispersed-nickel conductive coating for plastic enclosures that can be applied with conventional spray equipment and featuring surface resistivity of one ohm per square with coverage of 270 square feet per gallon, has been announced by Metex Corporation, Electronic Shielding Group, Edison, N.J.

Effective EMI/RFI shielding can be quickly and easily achieved on a wide variety of surfaces: acrylics, such as Plexiglass and Lucite; polycarbonates, such as Lexan; ABS; polyphyenylene oxide, such as Noryl; fiberglass, such as Hexcel F-185-1; sandblasted steel; as well as a broad range of other plastics and metals.

The air-cured, sprayable shielding liquid offers both excellent thermal shock and a high order of environ-



mental resistances. Test samples of the materials listed above have been successfully subjected to 5 cycles of thermal shock per MIL-STD-202E, Method 107D, Condition A and a 48 hour Salt Spray Test per MIL-STD-202E, Method 101D, Condition B.

Contact Metex Corporation, Electronic Shielding Group, 970 New Durham Road, Edison, N.J. 08817. Circle INFO/CARD #140.

Low Pass Filter

Sage Laboratories, Inc. introduces a new low pass filter which combines 10 KW peak power capability with freedom from spurious responses to above 10 GHz. The model FF2573 also features VSWR of less than 1.2/1 and an insertion loss of less than 0.3 dB over the passband from 962-to-1213 MHz.

Its minimum rejection is 30 dB at 1600 MHz and 45 dB through the eighth harmonic.

Contact Sage Laboratories, Inc., 3 Huron Dr., Natick, Mass. 01760. Circle INFO/CARD #120. □

Advertiser Index

1,000 MHz in one band —with IEEE-488 Bus.

S-D's 1702 Signal Generator.

Since many of today's test requirements go well beyond 520 MHz, Systron-Donner offers you the Model 1702 AM/FM Programmable Signal Generator—with 100 Hz resolution! The Model 1702 not only provides wide coverage but does it with the stability and accuracy of a synthesizer. Furthermore, the 1702's typical level flatness of ±1 dB is significantly better than other signal generators costing much more. The 1702 is one of a growing list of S-D instruments offering the IEEE-488 Bus interface.

New: Model 1618 Frequency Synthesizer. This 18 GHz Frequency Synthesizer is a state-of-the-art frequency source for the most demanding applications requiring high stability and accuracy. Using S-D's unique phase lock techniques, Model 1618 provides wideband coverage without the use of Output multipliers. IEEE-488 Bus interface is standard.

For details, contact Scientific Devices or Systron-Donner at 2727 Systron Drive, Concord, CA 94518.

S-D: The signal generator people.
And a lot more.

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