January/February 1984

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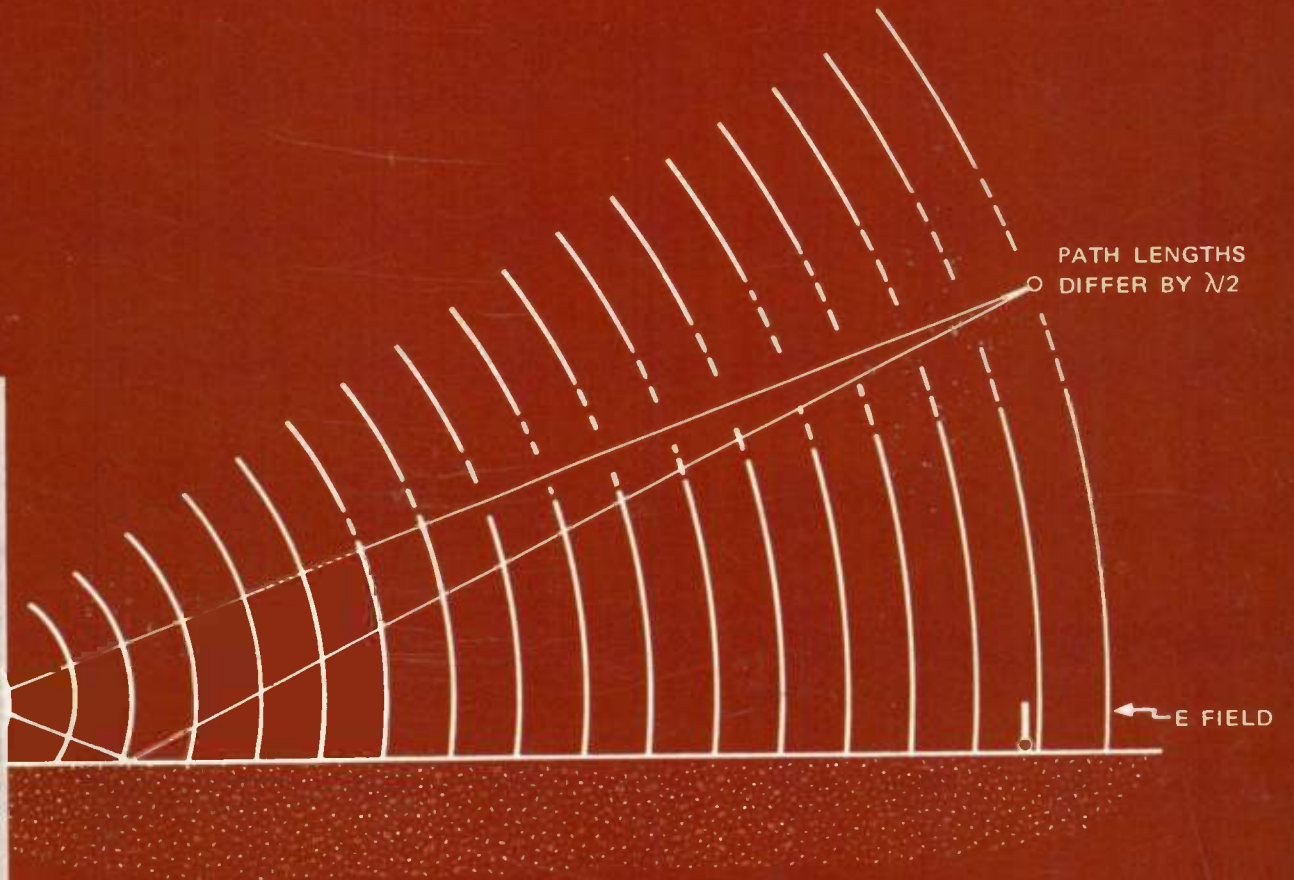
rf design



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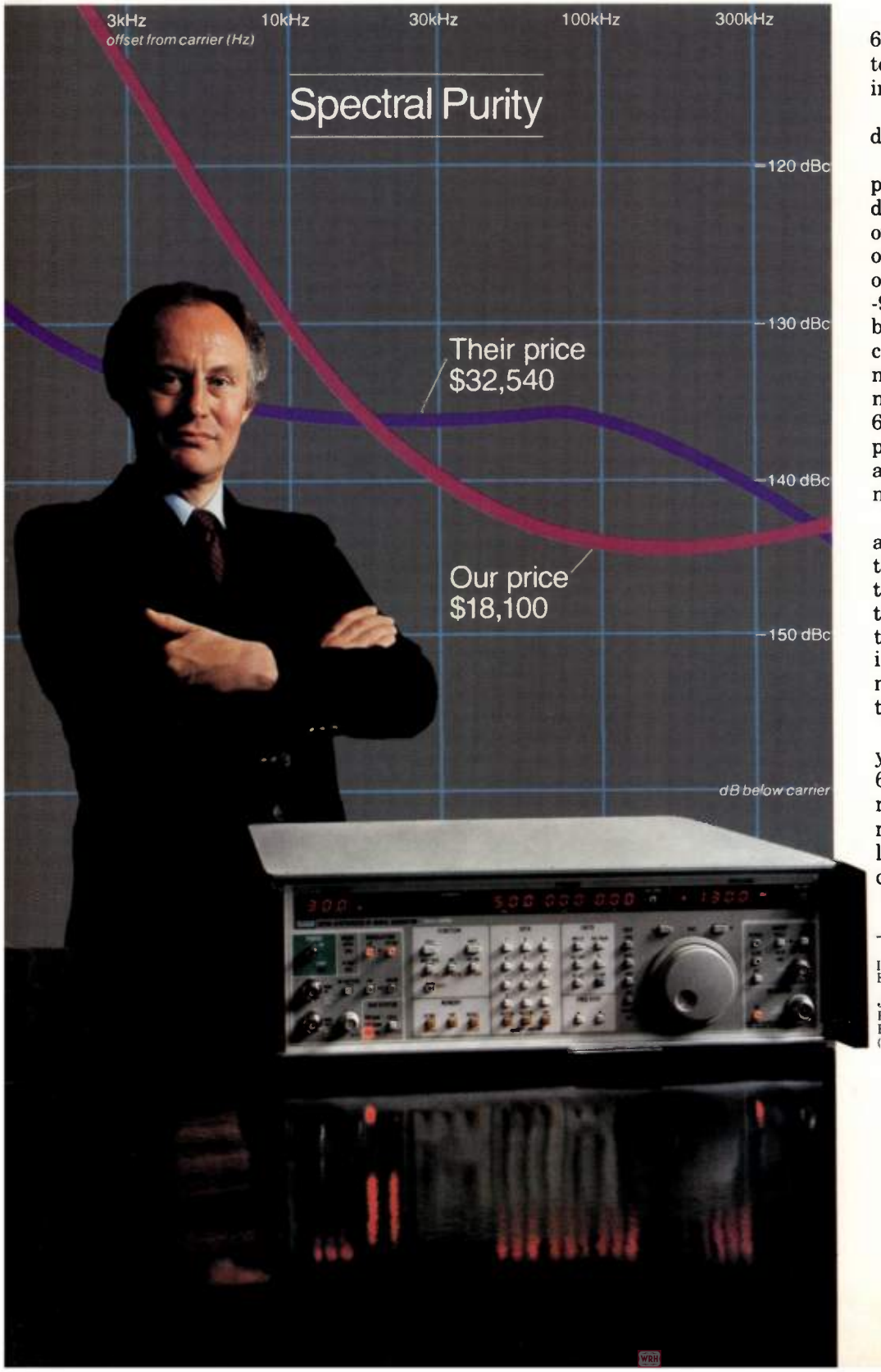

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r.f. design

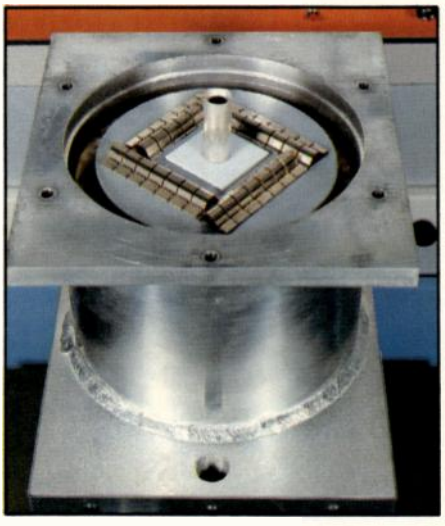
January/February 1984

Cover

1 **January/February Cover** — The front cover for this issue draws attention to the feature article concluding in this issue. In this article the authors have developed a set of basic formulas for the characteristics of monopole and dipole antennas. Part II of this article begins on page 12 of this issue.

Features

Departments



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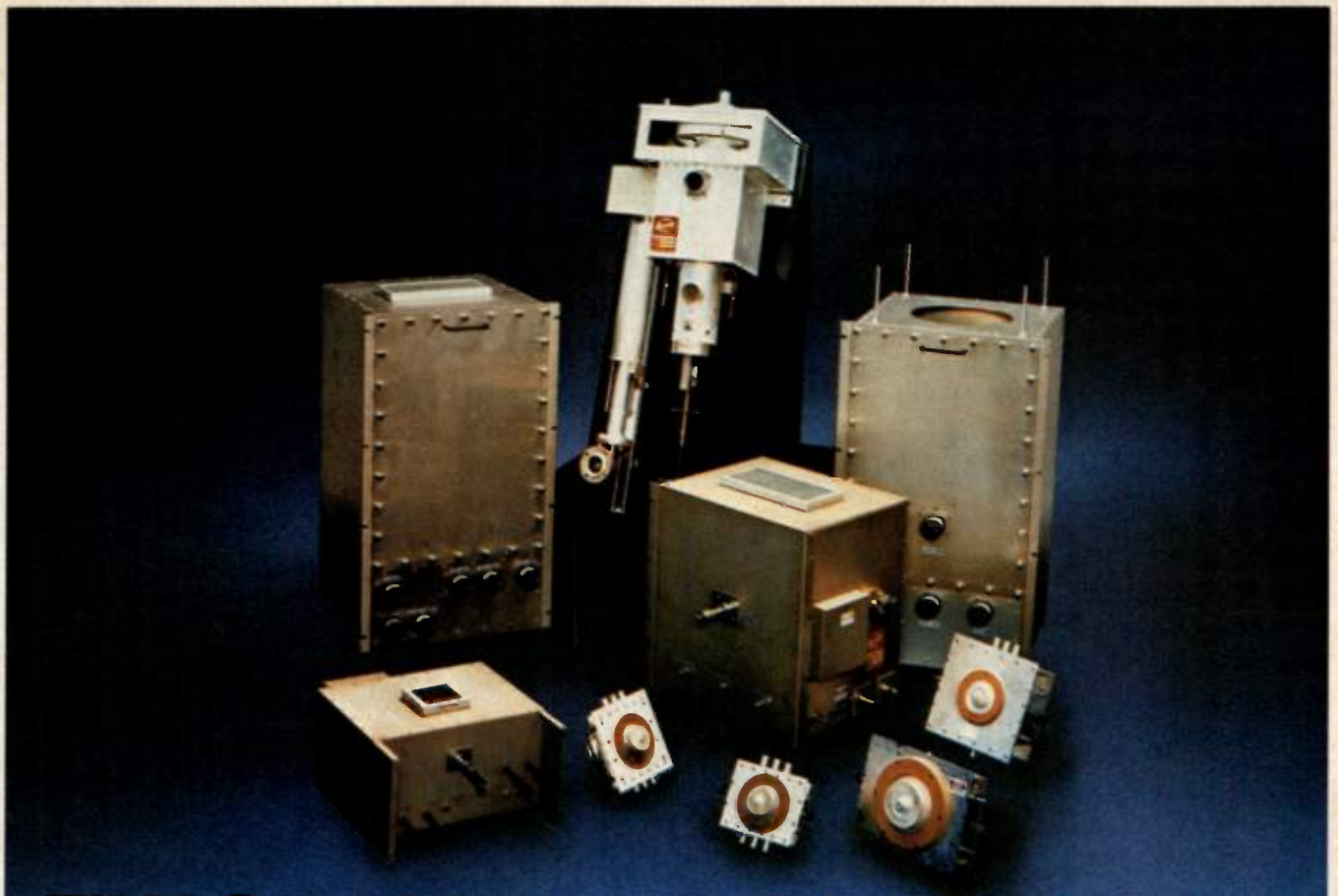

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CV-2250	3CX10,000U7	170–227	10 kW†
CV-2400	8874	420–450	300/1250 W*
CV-2800	3CX400U7	850–970	225 W
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*pulsed power
†peak sync, or 2.5 kW combined in translator service

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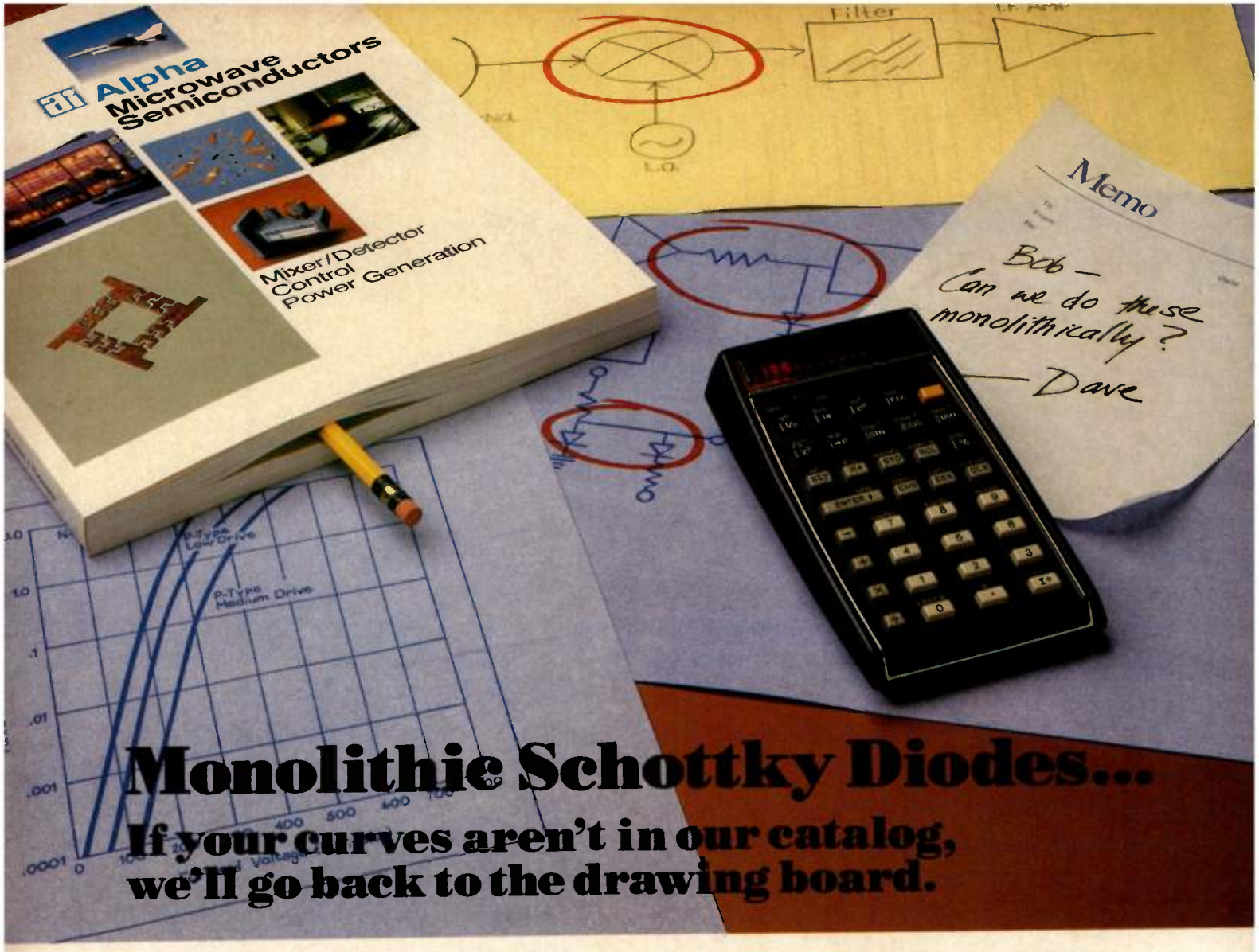

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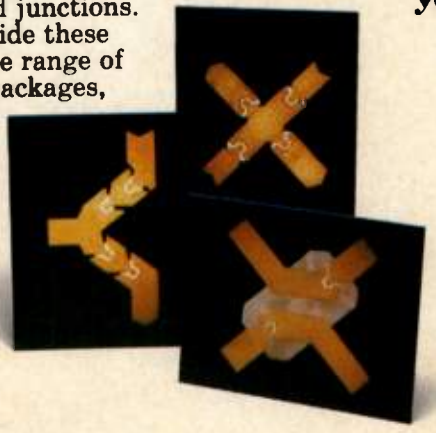

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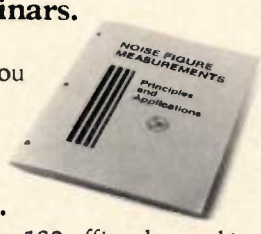

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INFO/CARD 6

Typos in Recent Article

Dear Sir,

I have just read Marvin Kefer's article "An Easy Method for Measuring Unloaded Q (Qu) for an Inductor or Filter Tank" in the March/April 1983 issue of *r.f. design*. I enjoyed the article very much and feel it will benefit me as soon as I get the time to construct the necessary test fixture. My intention is to use it in conjunction with a network analyzer.

As valuable as the article is, may I point out what I believe to be a few errors in the text.

1. Table 1, page 64. I believe R_{eff} (in the circuit drawing) should be R_c in both cases.

2. Item 3, page 64, $R = Q_u X_L$ probably should read $R_L = Q_u X_L$.

3. Item 4, page 64

$$R_c = \frac{R_{eff} R_L}{2(R_L - R_{eff})} \text{ should read } R_c = \frac{2 R_{eff} R_L}{R_L - R_{eff}}$$

$$R_{eff} = \frac{(R)(2R_c)}{2R + R_L} \text{ should read } R_{eff} = \frac{R_c R_L}{R_c + 2R_L}$$

4. Item 8 page 65. This equation is not familiar to me and, in any case, appears to be incorrect. It probably should read:

$$L = \frac{1}{\omega^2_0 (3.96 + C_c)}$$

The results given as 73.2 μH probably should read 73.2ηH.

5. Appendix III page 67. The denominator of the equation following EQ(9) reads $Q X'_{par}$ should probably read

$$\frac{1}{Q X_{par}}$$

6. Appendix IV page 68. This last comment is probably a bit nit picking and involves the derivation of the inductance error. The derived equation is of course, correct as shown. However, the derivation seems more complex than necessary in that it involves using the product rule, multiplying both numerator and denominator by dC and then substituting ω_o^2 for LC. A simpler way might be:

$$LC = \omega_o^{-2}$$

$$L = \omega_o^{-2} C^{-1}$$

$$\frac{dL}{dC} = -\omega_o^{-2} C^{-2}$$

$$dL = -\frac{1}{(\omega_o C)^2} dC$$

Very Truly Yours,
Dan O'Neill

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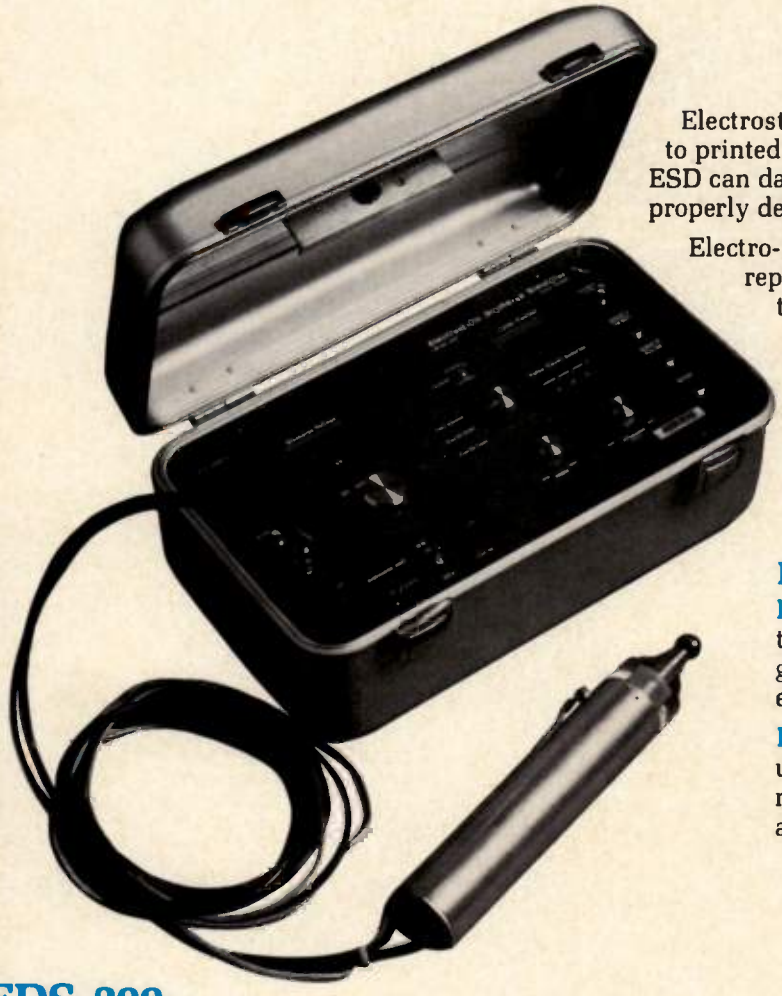

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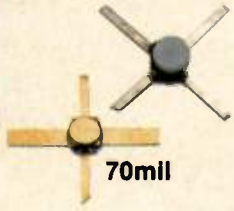

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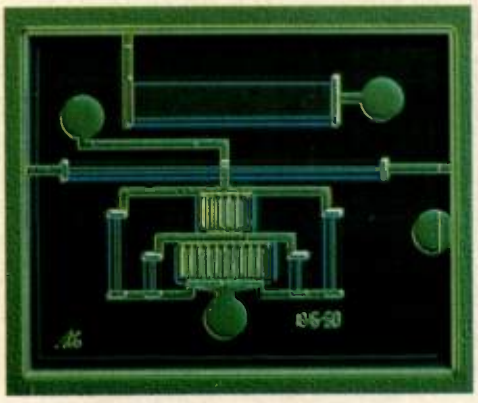

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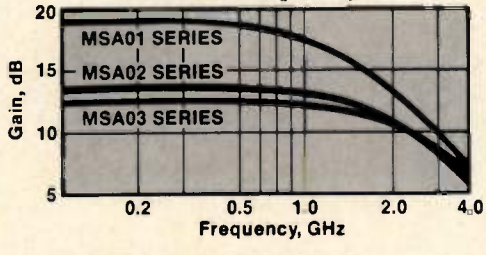

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Part II

GAIN, CAPTURE AREA, AND TRANSMISSION LOSS FOR GROUNDED MONOPOLES AND ELEVATED DIPOLES

Part II continues the development of basic formulas for the gain, capture area and other parameters of dipoles and monopoles by extending treatment to sky-paths and other useful variations.

By John W. Ames
William A. Edson
SRI International
Menlo Park, CA 94025

Part I of this article which appears in the November/December issue of *r.f. design* develops basic formulas for the gain, capture area, and other parameters of dipoles and monopoles. The following pages complete the treatment by extending it to sky-wave paths and other useful variations.

Sky Waves

Figure 5 introduces a break in the ground plan, so that pure surface waves no longer can travel from transmitter to receiver. The only coupling is provided by an elevated reflector, which corresponds to refraction in the ionosphere, or to a direct space path in a circumstance in which two isolated ground plans are multually visible.

Although the lossy section of ground suppresses the surface wave and any reflections from the surface, the ground in the immediate vicinity of the monopole remains intact, and causes local doubling of the electric field. The appropriate gains for the two forms of receiving antenna are, therefore, 3 for the monopole and 1.5 for the dipole. As before, we assume that the reflector (ionosphere) is sufficiently far above the ground so that the surface wave does not terminate on it, but that it is low enough so that d is a good approximation to the path distance, and the maxima of the antenna patterns are applicable. A straightforward application of Eq. (4) with the gain values from Table 1 gives a normalized received power of 9 for the monopole-to-monopole case and 4.5 for monopole to dipole. For these particular circumstances, and unlike the example in Figure 3, the effective capture area of the monopole is double (not half) that of the dipole.

It is important to note that relections to the dipole from the surface have been excluded, which may or may not correspond to a particular installation.

If the power received on a monopole is used to compute incident power density, by Eq. (10), some means must be employed to determine whether surface or space waves are being received. In the surface-wave case, the appropriate capture area is 0.06 λ^2, and the resulting power density for the incident wave and the field in the immediate vicinity of the monopole are equal. In the space-wave case, the gain and capture area are four times greater, 0.24 λ^2, and the resulting power density applies to the incident space wave, and not to the field at the monopole.

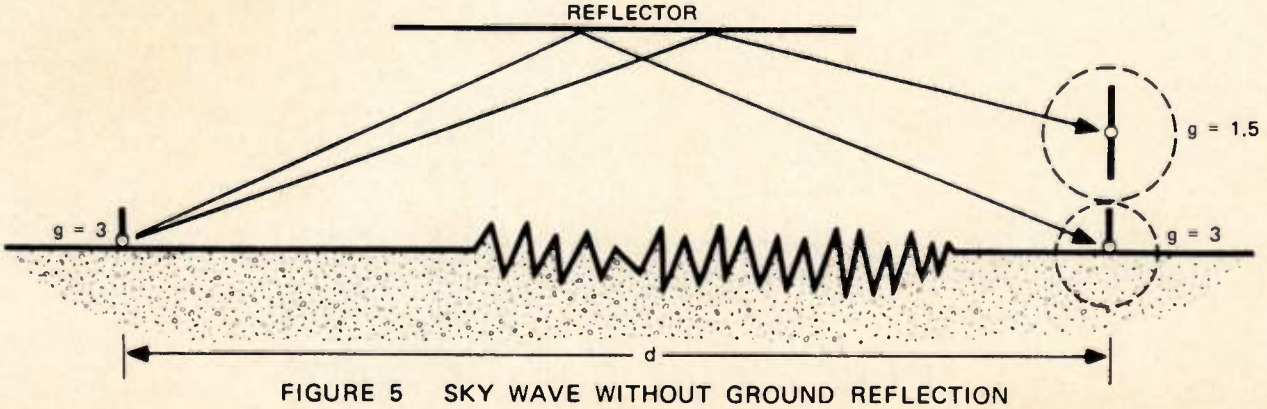

FIGURE 5 SKY WAVE WITHOUT GROUND REFLECTION

The self-consistency of this conclusion can be verified by noting from Table 2 that the power received on a monopole via space wave, Figure 5, is four times greater than that received via surface wave, Figure 3. Because the capture area applicable to the space-wave case is four times larger than that for the surface-wave cases, it follows that the power density inferred by dividing received power by capture area, Eq. (10), would be equal in the two cases, as it must be by Eq. (11).

In many applications, a dipole or one of its derivatives, such as a yagi or a log-periodic antenna, is mounted to take advantage of reflection from nearby ground (Figure 6). Again considering incident sky waves, and assuming that pattern maxima apply, we see that two waves reach the dipole and add in phase to double the voltage, which quadruples the received power. Thus, the effective gain of such a dipole is four times that of a dipole in free space, or 6 (7.78 dB) for short dipoles. This is the value that a manufacturer would normally advertise. The corresponding normalized received power, in Table 2, is 18. Note that there is no corresponding local-reflection increase for a monopole at the receiving location because use of the reflection from the immediate vicinity is inherent in the operation of the basic monopole. That is, if there were no ground immediately around the monopole, there would be no monopole.

The capture area for the combination of dipole and ground enclosed by the dashed line in Figure 6 may be computed by using Eq. (13); it is equal to 0.48 λ^2. The power density that results from dividing available power by capture area (Eq. 10) is, therefore, that which would exist in the vicinity of the antenna in the absence of the ground; in other words the power density of a single ray reaching the combined antenna/ground. The power density at the dipole itself is four times greater because of the reflection, but the capture area of the dipole alone is only 0.12 λ^2.

The results tabulated so far are self-consistent, but require care in making comparisons. The power density versus range (Eq. 11) for a single-mode space wave is the same as for a single-mode surface wave (assuming both are transmitted by a monopole). A dipole that receives only a space wave, as in Figure 5, will, therefore, capture the same power as a dipole receiving a surface wave, Figure 3 (4.5 in Table 2).

If we refer back to Figure 6 and imagine a monopole located just below the dipole, we can say that the monopole and the dipole experience the same power density; therefore, the 18-to-9 ratio of received power is consistent with Eqs. (23) and (24). To confirm that Figure 6 represents a locally uniform power density (in height) while Figure 5 does not, we must recognize that in both cases the monopole is, in effect immersed in both direct and reflected waves. In Figure 5, the wave that would reflect up to the dipole has been suppressed. As discussed earlier, the "reflected" wave at the monopole is incorporated in the basic monopole formulas, and cannot be suppressed.

Replacing the monopole transmitter in Figure 6 with a dipole leads to the configuration in Figure 7. Again the surface and direct waves are suppressed, but local ground reflections are included. In these circumstances, antennas are normally designed to make use of the reflection from nearby ground, and the gain is usually specified in terms of an elevation pattern, the maximum of which assumes the existence of a perfect local ground reflection. To correctly compute the gain of the antennas as installed, the four separate ray paths must be considered. First, a ray can go directly from the transmitting dipole to the elevated reflector and then directly to the receiving dipole. Another ray can reflect off the ground in front of the transmitting dipole, then proceed to the receiving dipole via the reflector. A third ray can be the converse of the second, while the fourth ray can involve a ground reflection at both ends. If we assume that the paths add in phase, which is equivalent to requiring that the gross path be at the angle of the pattern maximum of each antenna as installed, the received voltage is four times the received voltage for the single-ray case, and the power will be 16 times as great, for a normalized received power of 36. The effective gain of each installed dipole is therefore 6, as given in Table 1, and the basic transmission loss is given by Eq. (4).

More complex combinations of paths than those discussed here should be handled as separate paths rather than attempting to deduce effective antenna gains. For example, transmission between two monopoles via both ground and sky wave leads to a normalized received power of

$$\frac{W_r}{W_t}\left(\frac{4\pi d}{\lambda}\right)^2 = \left(\sqrt{\frac{3\times 3}{4}} + \sqrt{3\times 3}\right)^2 = 20.25 \qquad (28)$$

Use of Barrick's Transmission Loss

Barrick [1971] has derived corrections to the basic transmission loss for surface-wave propagation over the ocean, including effects of earth curvature and various degrees of ocean roughness. His results are referred to the expression

$$\ell_b = \left(\frac{2\pi d}{\lambda}\right)^2 = 1/4\left(\frac{4\pi d}{\lambda}\right)^2 , \qquad (29)$$

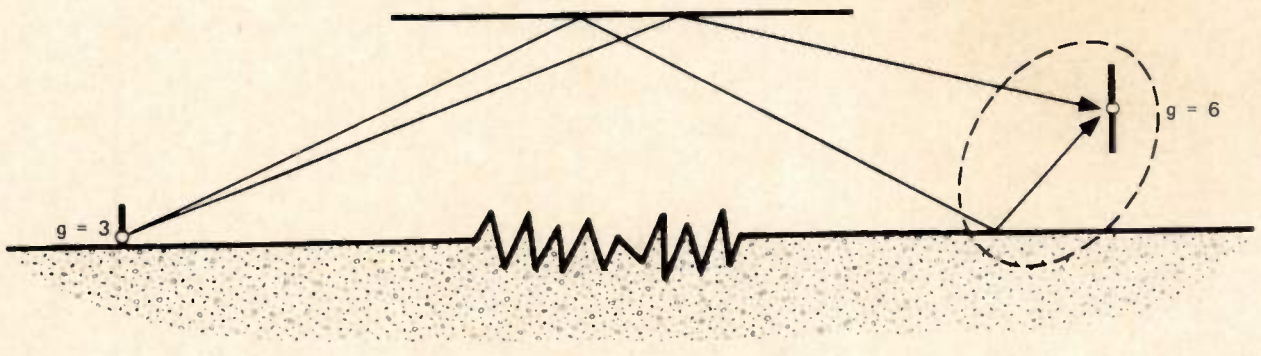

FIGURE 6 SKY WAVE WITH ONE LOCAL GROUND REFLECTION

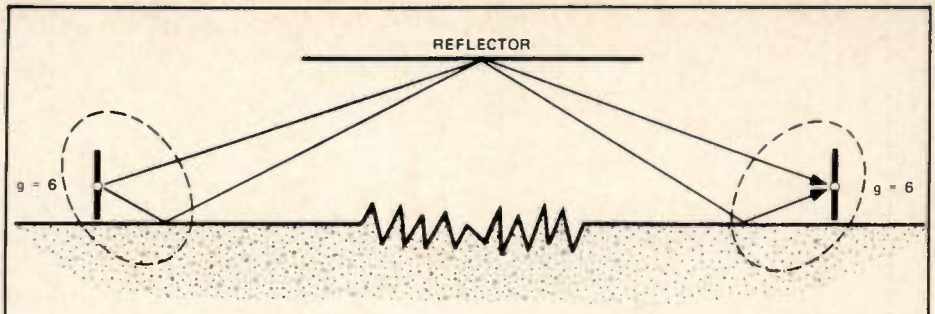

FIGURE 7 SKY WAVE WITH LOCAL GROUND REFLECTIONS AT EACH END

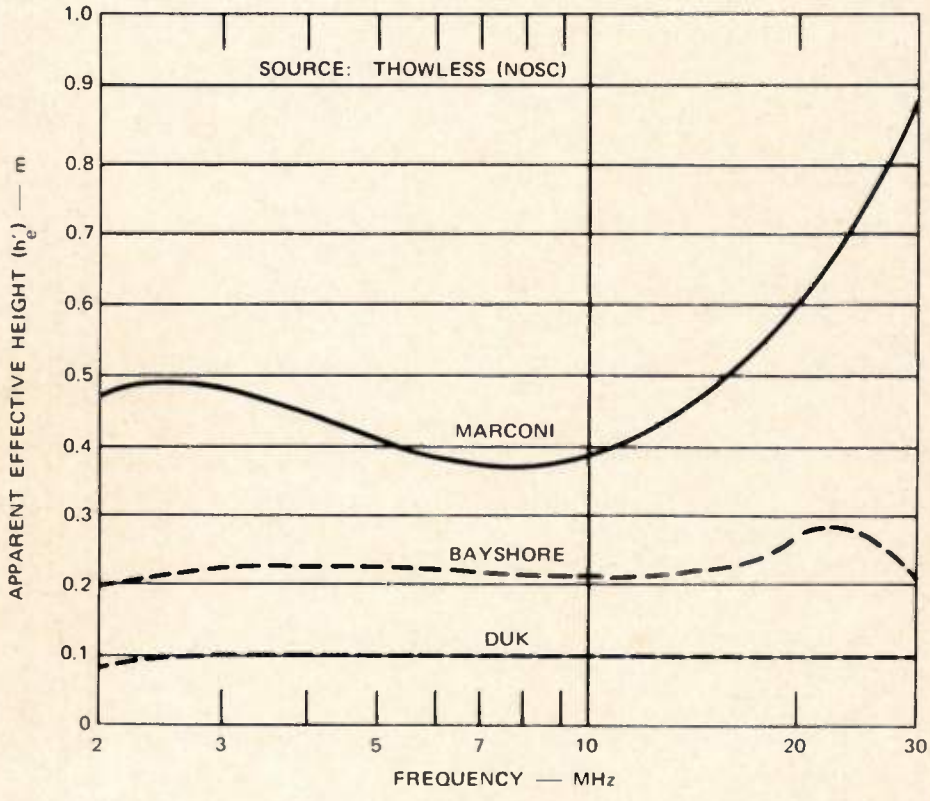

FIGURE 8 MEASURED APPARENT EFFECTIVE HEIGHT vs FREQUENCY

Errata

In part 1 in the last issue, the following errors were noticed. The corrected equations should be:

$$\frac{W_r}{W_t} = \frac{g_t g_r}{\ell_b} = g_t g_r \left(\frac{\lambda}{4\pi d}\right)^2 \qquad (8)$$

$$\frac{W_r}{W_t} = \frac{3}{4\pi d^2} \times \frac{3\lambda^2}{4\,4\pi} = \frac{3 \times 3}{4} \left(\frac{\lambda}{4\pi d}\right)^2 \qquad (22)$$

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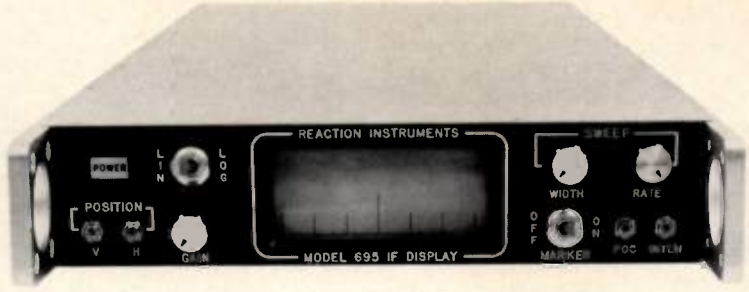

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which is correct for two isotropic antennas coupled by a direct wave and one perfect reflection — similar to our Figure 2. This procedure leads to a requirement to describe all antennas "as though they were in free space," even monopoles that have no existence without their ground plane.

We have used Eq. (6) consistently for path loss and to deduce values of antenna gain that are consistent with this usage. We may, therefore, use *Barrick's* [1971] results simply by adding 6 dB to his curves whenever he reports a value for transmission loss itself, and by using his *corrections* to transmission loss directly. The resulting value of basic transmission loss may then be used with the values of antenna gain given in Table 1.

Radiation Resistance

The concept of radiation resistance is required for many calculations of antenna performance. *Jordan* [Secs. 10.03 and 10.04 (1950)] gives a good discussion of radiation resistance. The power radiated by a current element of length, ℓ with uniform rms current, I, at a wavelength, λ, is given in Eq. 10-53 in *Jordan* [1950] as

$$P = 80\pi^2\left(\frac{\ell}{\lambda}\right)^2 I^2 \quad . \tag{30}$$

The coefficent of I^2 that yields power is conveniently identified as a resistance, specifically the radiation resistance R defined by

$$R\left(\frac{\text{current}}{\text{element}}\right) = 80\pi^2\left(\frac{\ell}{\lambda}\right)^2 \quad . \tag{31}$$

The application of the concept of radiation resistance to short real antennas, i.e., those having triangular rather than uniform current distribution, is succinctly described by *Jordan* [1950] as follows:

> "For the same current I (at the terminals) the (short) practical dipole of length ℓ will radiate only one-quarter as much power as the current element of the same length, which has the current I throughout its entire length. (The field intensities at every point are reduced to one-half, and the power density will be reduced to one-quarter.) Therefore, the radiation resistance of a practical short dipole is one-quarter that of the current element of the same length. That is

$$R\,(\text{short dipole}) = 20\pi^2\left(\frac{\ell}{\lambda}\right)^2 \quad . \tag{32}$$

> The monopole of height h, or a short vertical antenna mounted on a reflecting plane, produces the same field intensities above the plane as does the dipole of length $\ell = 2h$ when both are fed with the same current. However, the short vertical antenna radiates only through the hemispherical surface above the plane, so its radiated power is only one-half that of the corresponding dipole. Therefore, the radiation resistance of the monopole of height $h = \ell/2$ is

$$R\,(\text{monopole}) = 10\pi^2\left(\frac{\ell}{\lambda}\right)^2 = 40\pi^2\left(\frac{h}{\lambda}\right)^2 \quad . \tag{33}$$

> These formulas hold strictly for very short antennas only, but they are good approximations for dipoles of lengths up to one quarter wavelength, and monopoles of heights up to one-eighth wavelength."

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V_{DSX}	Drain to Source Voltage	V			10
V_{G1S}	Gate 1 to Source Voltage	V			-4.5
V_{G2S}	Gate 2 to Source Voltage	V			-4.5
I_D	Drain Current	mA			80
P_T	Total Power Dissipation	mW			200
$\lvert Y_{fs} \rvert$	Forward Transfer Admittance at V_{DS} = 5V, V_{G2S} = 0, I_D = 10 mA, f = 1.0 kHz	mS	20	30	
C_{iss}	Input Capacitance at V_{DS} = 5V, V_{G2S} = 0, I_D = 10 mA, f = 1 MHz	pF	1.5	2.0	2.5
C_{rss}	Reverse Transfer Capacitance at V_{DS} = 5V, V_{G2S} = 0, I_D = 10 mA, f = 1 MHz	pF		0.03	0.04
Gps	Power Gain at V_{DS} = 5V, V_{G2S} = 0, I_D = 10 mA, f = 900 MHz	dB	16	20	
NF	Noise figure at V_{DS} = 5V, V_{G2S} = 0, I_D = 10 mA, f = 900 MHz	dB		1.3	2.5

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The preceding derivation is important because occasionally Eq. (31) is encountered, quoted as the radiation resistance of a short dipole, without a prominent caveat that the current has been assumed to be uniform.

The triangular current distribution ceases to be a good approximation for dipoles longer than approximatley $\lambda/4$, or monopoles higher than $\lambda/8$. In such cases the sinusoidal current distribution is a good approximation. On this basis and through a lengthy calculation [*Jordan,* Sec. 10.06 (1950); *Ramo and Whinnery,* Sec, 21.07 (1955)], we obtain 73 Ω for the radiation resistance of a $\lambda/2$ dipole, and half that, or 36.5 Ω, for a grounded $\lambda/4$ monopole. (The approximate values given by Eqs. (32) and (33) would be 50 Ω and 25Ω respectively.)

Effective Height

Effective height, h_e, is a convenient quantity for relating the output of a receiving antenna to the field incident in its vicinity, i.e., the field that would exist in the absence of the antenna, but not the ground plane. It has physical credibility only for vertically polarized radiation, such as that characteristic of monopoles. Effective height is defined as the ratio of antenna open-circuit voltage to incident field strength. Its numerical value is derived [*Kraus* (1950)] by noting that

$$p = \frac{E_f^2}{120\pi} \quad , \tag{34}$$

and that the power available to a conjugate-matched load is

$$w_r = \frac{E_{oc}^2}{4R} \quad , \tag{35}$$

where

E_f = RMS field strength,
E_{oc} = antenna open circuit voltage,
R = radiation resistance.

Combining Eqs. (10), (34), and (35), we have

$$\frac{E_{oc}^2}{4R} = \frac{E_f^2 A}{120\pi}$$

or

$$h_e^2 = \frac{E_{oc}^2}{E_f^2} = \frac{4RA}{120\pi} \quad . \tag{36}$$

Using the values for R and A derived previously, we have, after some manipulation,

$$h_e \text{ (short dipole)} = \ell/2 \quad , \tag{37}$$

and

$$h_e \text{ (short monopole, surface wave)} = h/2 \quad . \tag{38}$$

These expressions are correct in all cases if evaluated using the total field at the antenna element. They are also correct for the arriving ray when no local reflection is involved, i.e., a dipole in space or a monopole receiving a surface wave. The larger values indicated by the fourth footnote in Table 1 relate the open-circuit voltage to the ray arriving in the vicinity of an antenna-ground combination, such as an elevated dipole or a monopole receiving a space wave.

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Application To Practical Antennas

The idealized antennas considered here are similar to many practical antennas, with the result that the gain values in Table 1 can be used directly. For example, inverted cone antennas are manufactured by both Granger Associates (Model 1794)* and TCI (Model 550)†; both are specified as having low-angle gain over perfect ground of approximately 5 dBi, which is equivalent to the 5.16 dB given in Table 1 for the quarter-wave monopole. The detailed gain-versus-elevation angle patterns given by the manufacturers, including any corrections for imperfect ground, would be used with Eq. (4) for all purposes except reception of surface waves, for which they would be reduced by 6 dB from the 0-degree perfect-ground value, that is, to approximately -1 dBi. (The latter assumes a good local grounding system, though not the good ground of infinite area necessary to maintain +5-dBi sky-wave gain at 0-degrees takeoff angle.)

Vertically polarized antennas consisting of arrays of dipoles provide still higher values of gain. For example, the near-horizon gain of the Granger Model 1703* antenna is 11 dBi and that of the TCI Model 503† is 12 dBi. For sky waves these antennas correspond to dipoles employing a ground reflection as in Figure 6. Their 3 or 4 dB of gain above the 8.17 dBi of a dipole over ground results from the array action. When employed for reception of surface waves, the ground reflection is absent, and, like any other antenna, these units would have 6 dB less gain than they do for sky waves. The proper gain value to use for the Granger or TCI antennas for surface-wave reception would, therefore, be 5 or 6 dBi. Note that although the antenna itself does not require a local ground, the strength of the surface wave itself depends on ground constants.

Horizontally polarized antennas are treated the same as vertically polarized dipoles, except that their gain is always a maximum at some angle above the horizon, and they neither launch nor receive surface waves. For example, the Granger Model 1765 broadband dipole, with a maximum gain of approximately 8 dB, corresponds to the $\lambda/2$ dipole-over-a-ground of Table 1.

*Granger Associates, 3101 Scott Boulevard, Santa Clara, CA 95051.

†TCI, 1625 Stierlin Road, Mountain View, CA 94045.

Active Receiving Antennas

Receiving antennas that combine a low-noise preamplifier with a short, vertical monopole antenna are called active receiving antennas and are useful in several applications. The effective gain of such antennas can be determined by measuring an apparent effective height and calculating a corresponding gain. A report by *Thowless* [1978] describes a measurement method and results for three commercially available antennas.

A known surface wave field, E_f, is generated on a flat conducting plane, and the resulting output open circuit voltage, E_{oc}, of the antenna system is measured. The ratio E_{oc}/E_f is the apparent effective height, h'_e, in meters. From Eqs. (23) and (36), we have

$$h'^{2}_{e(sfc)} = \frac{4 R g_{sfc} \lambda^2}{4\pi(120\pi)} \qquad (39)$$

and

$$g_{sfc} = \frac{120\pi^2 h'^{2}_{e(sfc)}}{R\lambda^2} \qquad (40)$$

or

$$G_{sfc} = -35.8 + 20 \log_{10}(h'_e f_{MHz}) \text{ dBi} \quad . \qquad (41)$$

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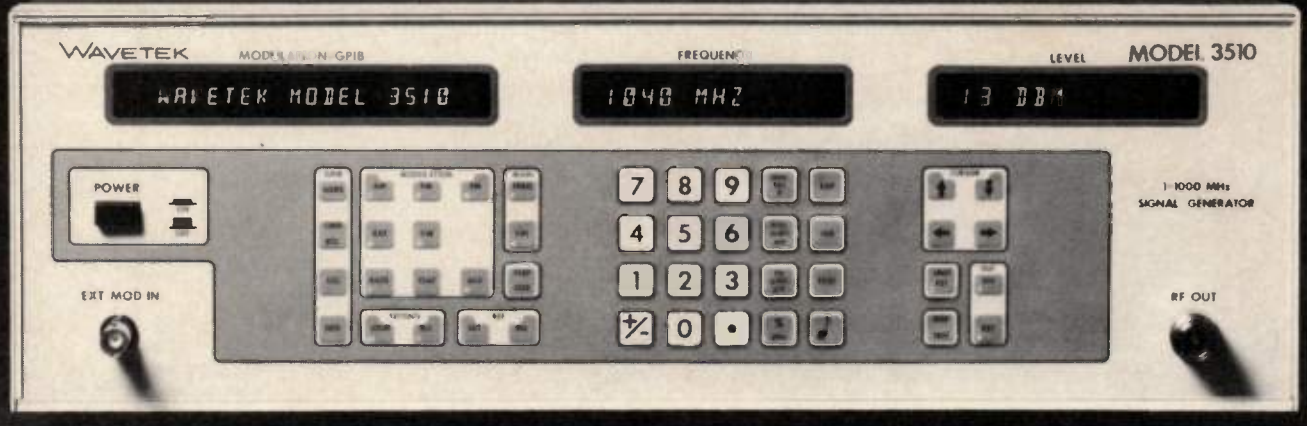

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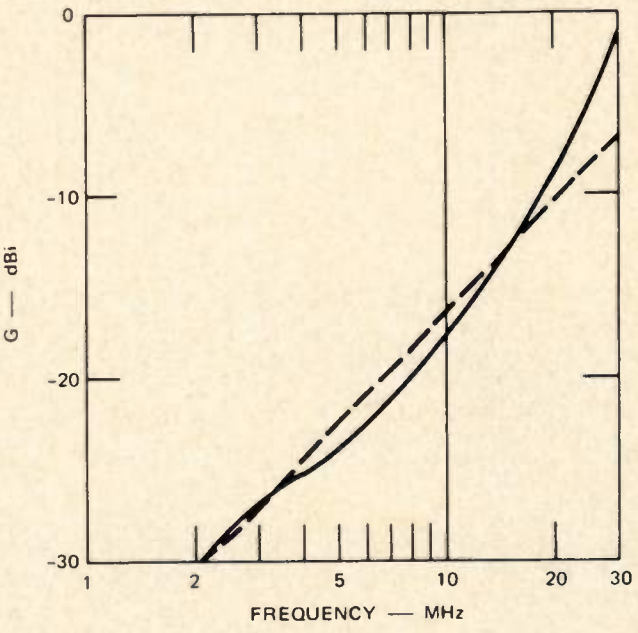
Figure 9.
Gain of Marconi Active Antenna



The notation h'_e follows that of *Thowless* [1978] to indicate an *apparent* effective height because the output voltage is a function of the amplifier parameters as well as of the antenna element itself. R is the nominal output impedance of the active antenna amplifier, normally 50Ω. The subscript (sfc) serves as a reminder that the measurement employs surface waves; therefore, the derived gain applies to reception of surface waves. As discussed earlier, the gain for sky waves would be

$$g = 4\, g_{sfc}$$

or

$$G = G_{sfc} + 6\,\text{dB} \quad . \tag{42}$$

Thowless [1978] has measured the apparent effective height of a Marconi Model H-33-5500-01 active antenna, with the results plotted in Figure 8. For the accuracy required in system modeling, 0.5 m is an adequate approximation to the apparent effective height. If we apply Eqs. (41) and (42), we obtain the gain (for sky waves) plotted in Figure 9, which is equivalent to, and within 1 or 2 dB of equaling, Figure 7 of *Thowless* [1978]. The dashed line shows the linear approximation

$$G(\text{dBi}) = -36 + 20 \log_{10} f_{\text{MHz}} \ , \tag{43}$$

which is useful for system modeling.

Conclusion

The concepts of basic transmission loss, antenna gain, power density, capture area, and effective height can be made compatible for practical calculations with surface and space waves, but at the expense of attibuting to a monopole 6 dB less gain when it is receiving surface waves than when it is receiving space waves or transmitting either space or surface waves.

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Methodology for the Design of Narrow Band, Low Noise Microwave Amplifiers

By Arturo Serrano Santoyo, Jose Luis Medina Monroy
CICESE Research Center, Baja California, Mexico

The technique presented in this article is based on the selection of a reflection coefficient Γ that provides the minimum noise factor as well as moderate gain for a single stage microwave amplifier. By means of a simple computational algorithm, it is possible to obtain the desired value of Γ for a given narrow frequency range. From this value of Γ, the input and output matching circuit elements of the active device can be calculated directly.

I. Introduction

There are basically two methods used in the design of a low noise microwave amplifier, network synthesis or trial and error using the Smith Chart. Network synthesis gives fast and efficient results when proper computer programs are used. Trial and error method is based on the selection of input and output reflection coefficients from which proper matching networks are calculated. One way to obtain such reflection coefficients is by using circles of constant gain and noise figure displayed in the Smith Chart. When the design of the amplifier is performed at several frequencies, this method becomes very tedious due to the large number of circles involved.

In the technique presented in this article, instead of using the circles to arrive at the proper reflection coefficients, a simple computer algorithm that provides the desired Γ is used. By means of this algorithm, graphs of noise figure and gain versus input reflection coefficient Γ_{oi}, are obtained for a particular frequency range.

II. Description of the Design Method

A basic block diagram of a microwave amplifier is shown in Fig. 1. The transistor used to describe the present technique is the GaAsFET HFET1101 whose noise and S parameters for the 3.7-4.2 GHz frequency band are shown in Table I.[1,2].

Fig. 2 shows the location of S_{11}^* and Γ_o for the HFET1101 in the inferior, central and superior frequencies of the band. The particular region in which the gain and noise figure graphs are going to be obtained are also shown.

As it is generally known, matching to S_{11}^* provides maximum unilateral transducer gain for the input circuit. Matching to Γ_o gives minimum noise figure. S_{11}^* and Γ_o for a particular device and frequency are in general different. This conflict has to be resolved depending on the requirements of the desired amplifier and on the properties of the active device and matching circuits. If a minimum noise amplifier is desired, the optimum input reflection coefficient Γ_{oi} will be equal to Γ_o and the output reflection coefficient Γ_∞ can be calculated by equation 1[3]

$$\Gamma_\infty = \left| S_{22} + \frac{S_{12}S_{21}\Gamma_{oi}}{1-S_{11}\Gamma_{oi}} \right|^* \qquad (1)$$

A compromise between maximum gain and minimum noise figure can be established considering the values of Γ_o and S_{11}^* for a particular device. In the case of the HFET1101, it can be observed from Fig. 2 that the variation of Γ_o towards S_{11}^* improves the gain, but at the same time the noise figure increases. On the other hand, variation of S_{11}^* towards Γ_o decreases the gain, but improves the noise figure value.

The methodology proposed for the design of a narrow band low noise microwave amplifier for a particular frequency range consists of the following steps:

1) Selection of the desired Γ_{oi}. To obtain the proper input reflection coefficient, a simple computer program performs three basic operations: i) Variation of reflection coefficient from Γ_o to S_{11}^* in a straight path, that is to say, varying simultaneously the magnitude and angle of Γ_{oi}. Fig. 3 shows results of this operation. ii) Variation of the angle of Γ_{oi} fixing the magnitude constant. Fig. 4 shows results of this operation having $|\Gamma_{oi}| = 0.6$ and varying the angle of Γ_{oi} from 70° to 130°. iii) Variation of $|\Gamma_{oi}|$ fixing the angle constant. Fig.5 shows results for this operation obtained with $\angle\Gamma_{oi} = 100°$ and varying $|\Gamma_{oi}|$ from 0.3 to 0.9

2) the output reflection coefficient Γ_∞ is calculated by equation 1.

3) Noise figure for each Γ_{oi} selected is obtained with equation 2 as follows[4]:

$$FR(dB) = 10 \log \left| Fmin + \frac{4r_n|\Gamma_{oi}-\Gamma_o|^2}{(1-|\Gamma_{oi}|^2)|1+\Gamma_o|^2} \right| \qquad (2)$$

where r_n is the equivalent normalized noise resistance, Fmin is the minimum noise figure.

4) Transducer gain for the amplifier stage is calculated by equation 3.[5]

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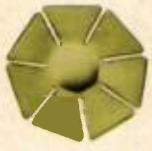

The Instrument Specialties' "Mushroom" is a two-piece assembly consisting of a beryllium copper top plate and plastic push rivet which mounts through a .250 inch diameter hole.

97-700

The "Pop-Up" is a spring-loaded brass assembly in a serrated cup for easy snap-in mounting. One pound of compression pressure yields over 1000 pounds contact force per square inch.

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Used with both the Mushroom and Pop-Up assemblies, this brass contact plate acts as a mating surface of consistent quality which mounts in the opposing plane.

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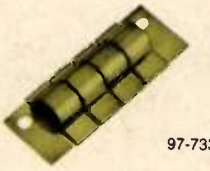

These beryllium copper grounding bars are mounted on chassis and cabinet frame components by riveting, heat staking, or by using common hardware such as screws and fasteners. Redundancy of contact points is attained by varying the number of fingers. Instrument Specialties also offers many custom variations.

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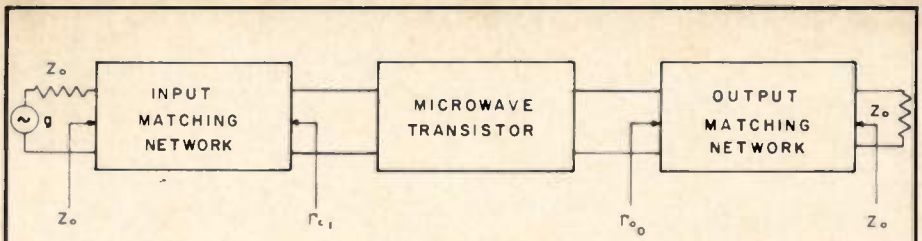

Figure 1. Microwave amplifier block diagram.

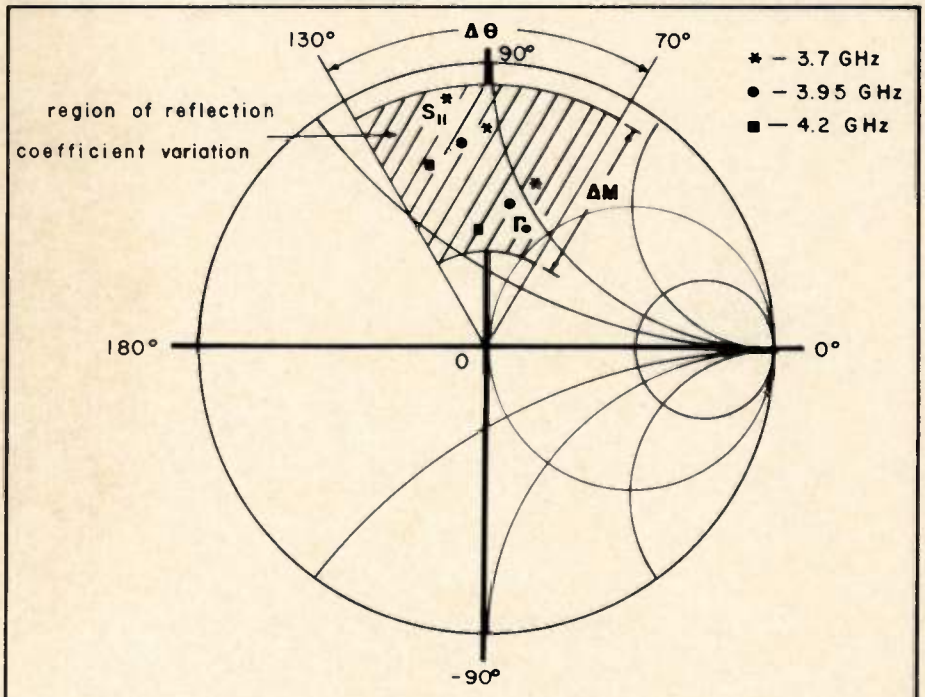

Figure 2. Smith chart showing Γ_o and S_{11}^* for the lowest, and highest frequencies (HFET1101).

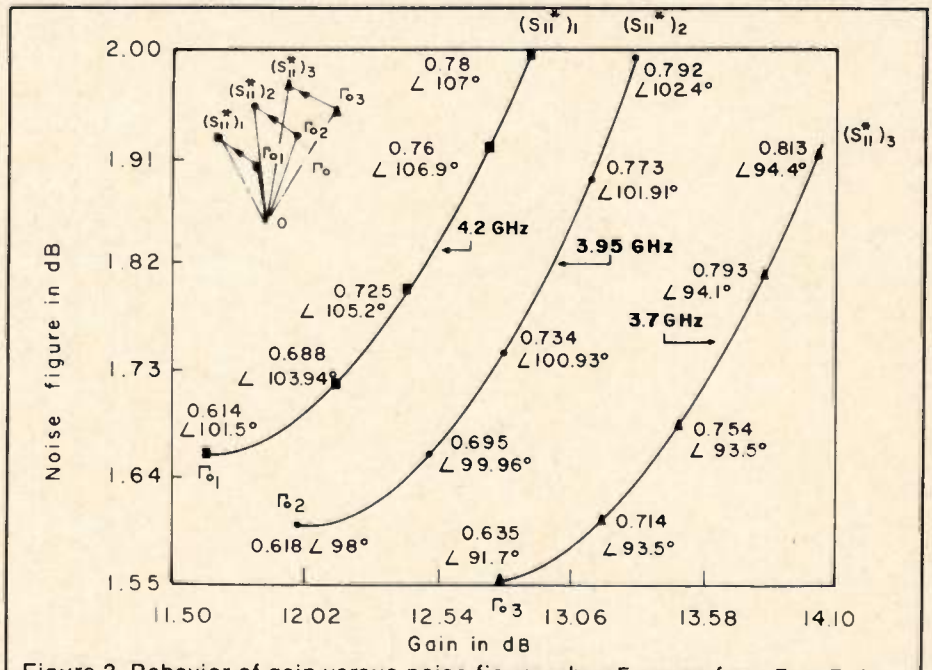

Figure 3. Behavior of gain versus noise figure when Γ_{oi} goes from $\Gamma_{oi} = \Gamma_o$ to $\Gamma_{oi} = S_{11}^*$ for the frequencies 3.7 (Δ), 3.95 (O) and 4.2 (□) GHz.

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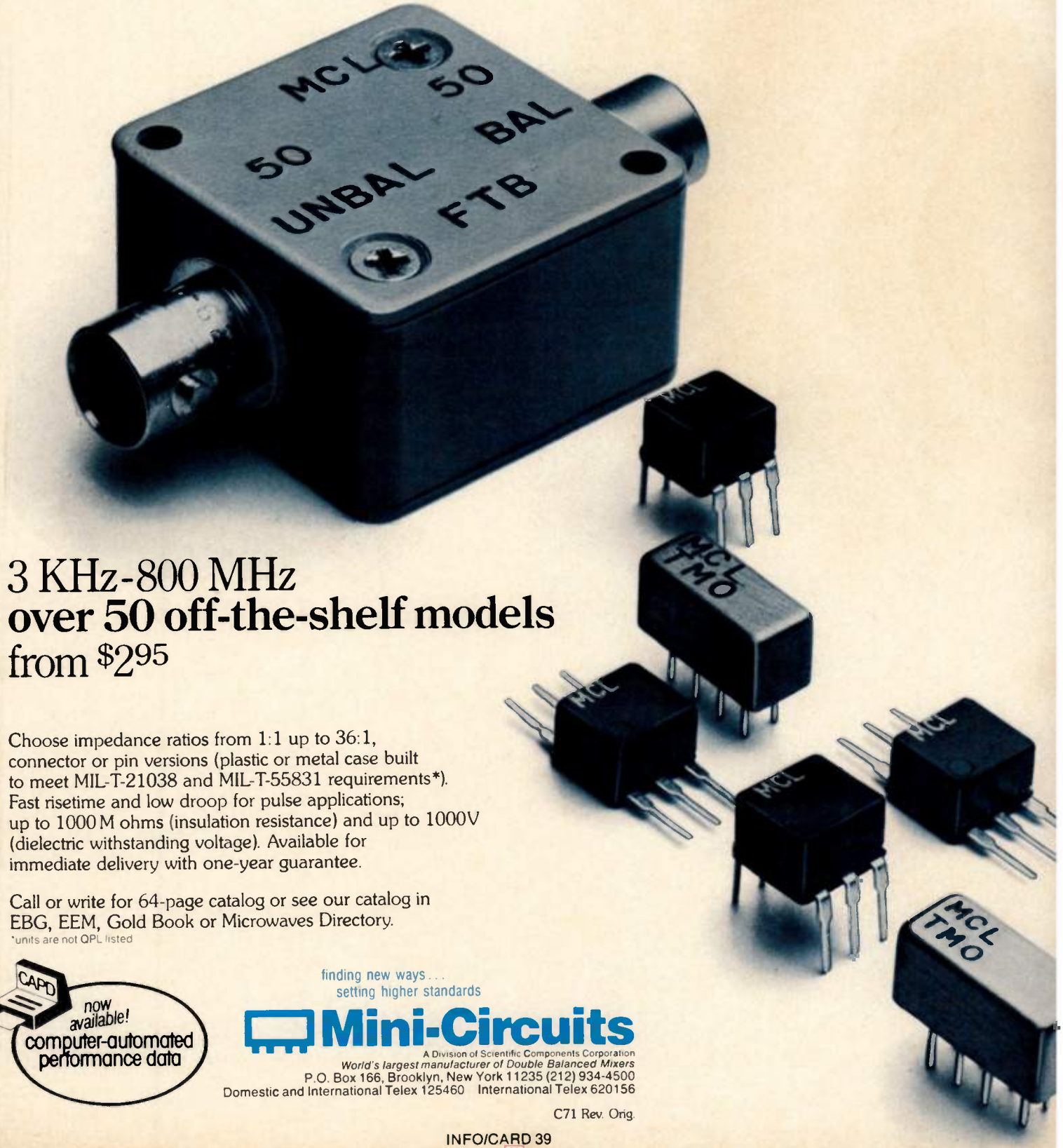

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$$GT(dB) = 10 \log \left| \frac{|S_{21}|^2(1-|\Gamma_{oi}|^2)(1-|\Gamma_{\infty}|^2)}{|(1-S_{11}\Gamma_{oi})(1-S_{22}\Gamma_{\infty}) - (S_{12}S_{21}\Gamma_{oi}\Gamma_{\infty})|^2} \right| \quad (3)$$

5) Having obtained Γ_{oi} and Γ_{∞} from the graphs shown in figs. 3, 4 and 5, the associated impedances are as follows[6]:

$$Z_{oi(o)} = \left| \frac{(1-|\Gamma_{oi(o)}|^2)Z_o}{1+|\Gamma_{oi(o)}|^2-2|\Gamma_{oi(o)}|\cos(\angle\Gamma_{oi(o)})} \right| + j \left| \frac{2Z_o|\Gamma_{oi(o)}|\sin(\angle\Gamma_{oi(o)})}{1+|\Gamma_{oi(o)}|^2-2|\Gamma_{oi(o)}\cos(\angle\Gamma_{oi(o)})} \right| \quad (4)$$

6) From the impedance values given by equation 4 the matching circuits are easily obtained. There are several methods widely described in the literature[7, 8, 9].

The behavior of gain and noise figure for each of the frequencies in the analysis can be observed from fig. 3. If a minimum noise amplifier is desired, it is convenient to select that Γ_{oi} corresponding to Γ_o at the highest frequency. This is done to compensate frequency roll-over effects of the transistor, If a minimum noise figure amplifier is not desired, or if a higher gain value is necessary, the selection of another Γ_{oi} can be easily don by means of fig. 3.

The graphs of figs. 4 and 5 show the behavior of gain and noise figure when the same Γ_{oi} is presented in the whole frequency range. This situation is not particularly true in practice because for each frequency there is a value of Γ_{oi}. However, for some narrowband designs the approximation can be acceptable.

III. Design Example

The requirements of the amplifier for the design example are:

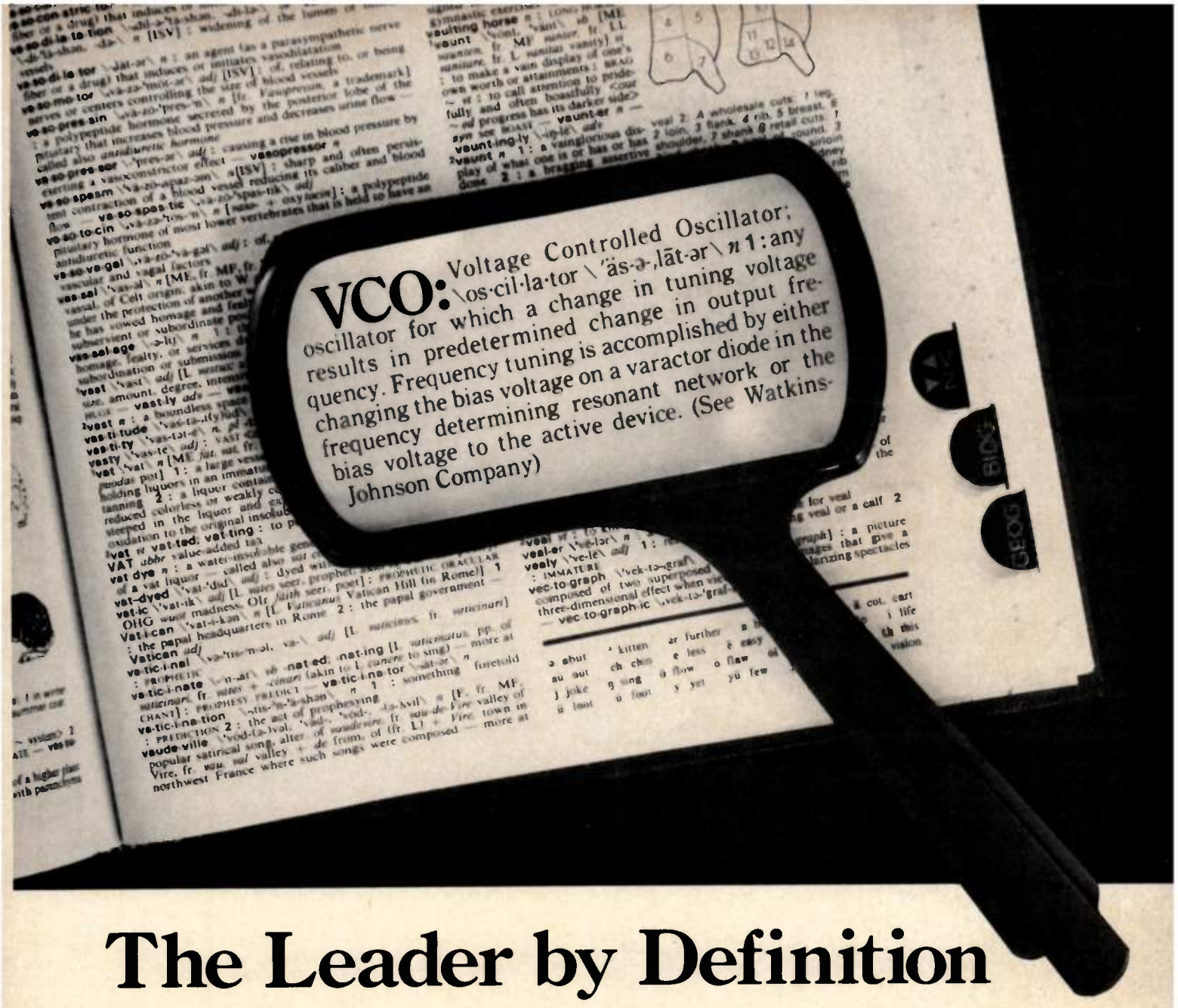
Maximum Noise Figure 1.8 dB
Minimum Gain 10 dB
Frequency Range (3.7-4.2) GHz.

The transistor used for the example is the HFET1101 whose parameters are given in Table I. Using the methodology proposed in this article, a Γ_{oi} is selected from Fig. 4. The specific Γ_{oi} selected is $\Gamma_{oi} = 0.6\angle -105°$. For this particular Γ_{oi}, we obtain $F_R = 1.64$ dB and $G = 12.29$ dB at the central frequency. At the highest and lowest frequency we observe that a noise figure $F_R = 1.68$ can be obtained with a gain flatness of around 1 dB in the whole frequency range. For the Γ_{oi} selected, Γ_{∞} is obtained by equation 1) giving $\Gamma_{\infty} = 0.69\angle 73.81°$.

The associated impedances are calculated from equation 4) giving:

$$Z_{oi} = 19.155 + j34.692 \text{ and } Z_{\infty} = 23.922 + j60.765$$

To calculate the matching networks, an in-house computer program was used[7]. This program gives the electrical length and impedance value of each one of the matching circuit elements. The values for Z and Θ are show in Table II. This same program has the capabilities of analyzing the behavior of the complete amplifier stage in the whole frequency range. Final results for this amplifier stage are shown in Table III. By further optimization[10] of the network elements, the general behavior of the amplifier can be improved. Tables



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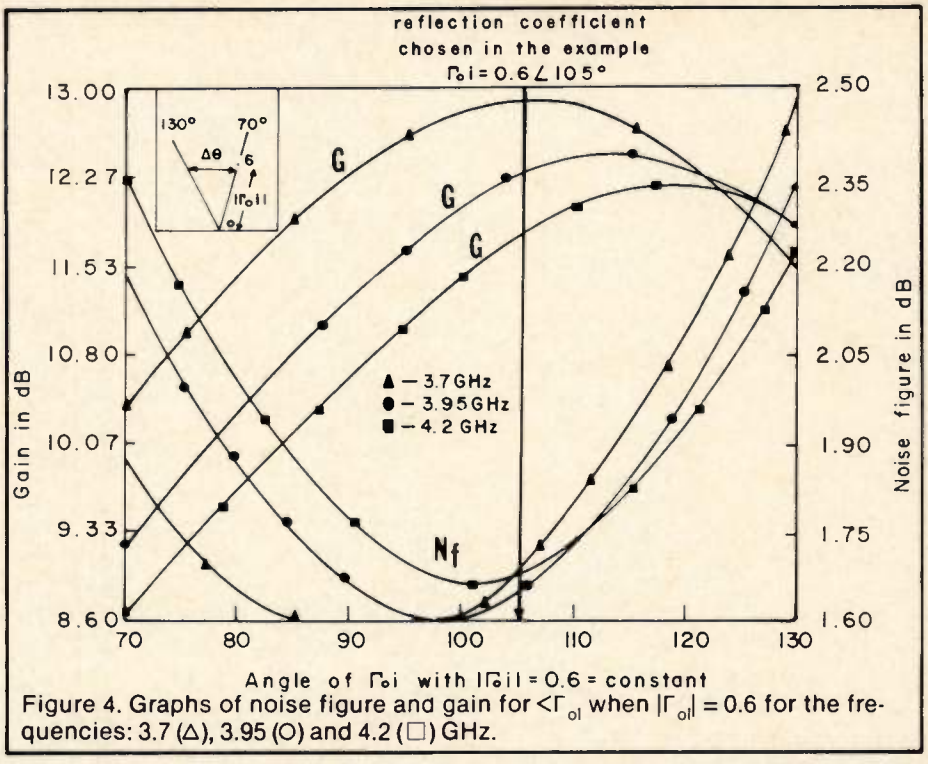

Figure 4. Graphs of noise figure and gain for $<\Gamma_{oi}$ when $|\Gamma_{oi}| = 0.6$ for the frequencies: 3.7 (Δ), 3.95 (O) and 4.2 (□) GHz.

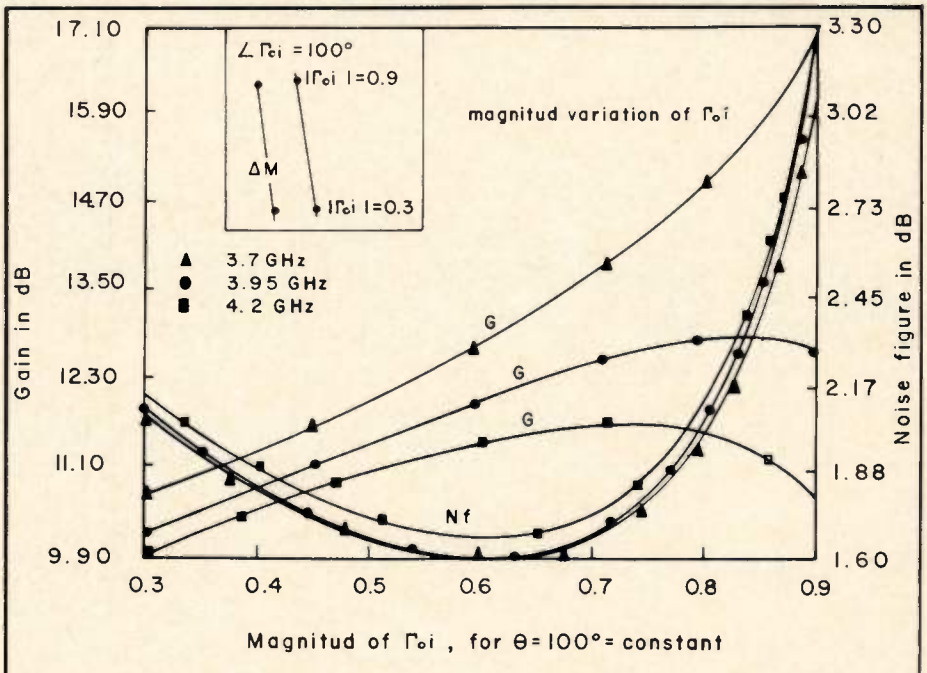

Figure 5. Variation of $|\Gamma_{oi}|$ with $<\Gamma_{oi} = 100°$ for the frequencies: 3.7 (Δ), 3.95 (O) and 4.2 (□) GHz.

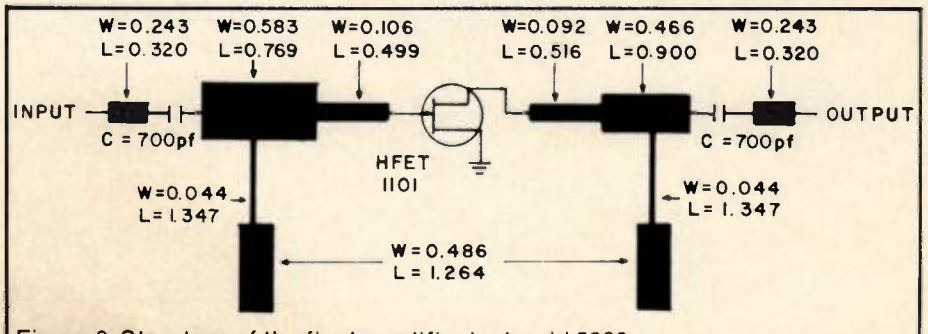

Figure 6. Structure of the final amplifier in duroid 5880.

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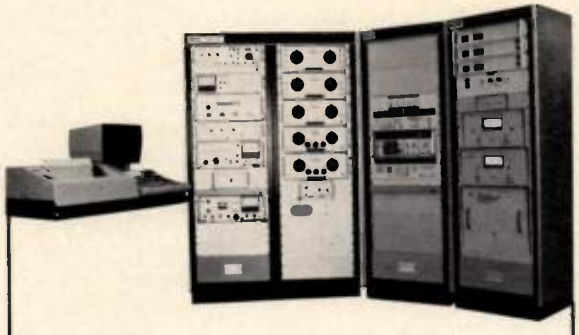

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FREQ. (GHz)	S_{11} (MAG ANG)	S_{21} (MAG ANG)	S_{12} (MAG ANG)	S_{22} (MAG ANG)	Fmin (dB)	Γ_o - (MAG ANG)	r_n
3.7	0.813 -94.4	1.990 88.3	0.067 29.6	0.067 -56.0	1.548	0.635 91.7	0.454
3.95	0.792 -102.4	1.995 81.1	0.070 26.0	0.569 -60.8	1.600	0.618 98.0	0.463
4.2	0.780 -107.0	1.936 76.9	0.071 23.7	0.653 -63.7	1.660	0.614 101.5	0.42

TABLE I

HFET1101 S AND NOISE PARAMETERS

IV and V show the final results of the designed amplifier. Fig. 6) shows the structure of the designed amplifier when DUROID 5880[11] substrate is used.

IV. Conclusions

The technique presented in this article provides an alternative design methodology. There are at the present several computer programs that help the high frequency electronics designer in his task. This one in particular, is a simple technique by which narrow band designs can be obtained with satisfactory results. Accuracy in the method depends on the bandwidth magnitude on the precision of the measured S parameters and on the quality of the matching networks. The analysis and the information provided by this methodology have been of great importance for the design of low noise TVRO microwave in our laboratory. □

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3	LT	25.000	53.158
4	LT	94.871	28.410
5	TR	HFET101	------
6	LT	95.000	34.737
7	LT	38.884	85.501
8	CS	700pf	------
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TABLE II
VALUES OF Z AND θ FOR THE DESIGN EXAMPLE

FREQ. (GHZ)	NF (dB)	G (dB)	VSWRI	VSWRO
3.70	1.825	11.638	6.14	2.22
3.95	1.640	12.059	3.64	1.026
4.20	1.671	10.889	3.74	1.734

TABLE III
GENERAL CHARACTERISTICS OF THE DESIGNED AMPLIFIER

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2	CS	700pf	------
3	LT	26.000	55.000
4	LT	81.000	34.000
5	TR	HFET1101	------
6	LT	87.000	35.000
7	LT	31.000	64.000
8	CS	700pf	------
9	LT	50.000	20.000

TABLE IV.
OPTIMIZED VALUES FOR THE MATCHING CIRCUIT ELEMENTS

FREQ. (GHz)	NF (dB)	G (dB)	VSWRI	VSWRO
3.70	1.764	10.841	5.06	3.22
3.95	1.620	11.426	3.61	1.78
4.20	1.700	10.816	4.09	1.13

TABLE V. OPTIMIZED RESULTS FOR THE DESIGNED AMPLIFIER

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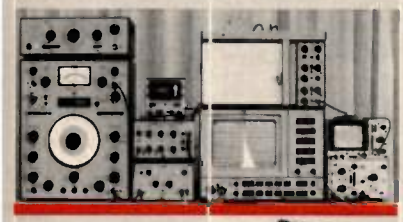

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Improved Testing for EMC Gasket Performance

A New Coaxial Test Fixture for Evaluating EMC Gasket Material Effectiveness

By William Stickney
Instrument Specialties Company, Inc.

Electromagnetic shields are used to minimize the transmission of electromagnetic energy across their boundary, either to protect sensitive circuitry from ambient interference, or to prevent unacceptable EMI emissions. Electromagnetic compatibility (EMC) gaskets maintain the continuity of a shield across joints and closures, and are thus commonly used around doors, access panels, ventilation openings, cable connectors, and many other components.

Typical gasket materials are metal finger stock, conductive elastomer, compressed wire mesh, elastomer filled screen, oriented wire mesh, and helical metal strip; each has its advantages. These materials are available in a variety of sizes and shapes, and some also function as environmental seals. Parameters used to select a proper material for an application include: shielding effectiveness, size, weight, closure force, corrosion resistance, ease of installation, fatigue life, and cost. Of all the gasket material characteristics, *effectiveness* is the most difficult to measure.

Two methods of evaluating EMC gasket material effectiveness are in use today: radiated energy measurement, and conducted energy measurement.

The radiated energy approach is the older of the two methods and is based on MIL-STD-285. In this procedure, gasket material is used to seal an opening in a shielded enclosure. A transmitting antenna is located outside of the enclosure, and a receiving antenna is located inside. Effectiveness is basically the difference in measured signal strength inside the enclosure before and after the gasket is installed. Poor repeatability resulting from enclosure resonance, reflections, and antenna loading, as well as a lack of standardization in mechanical dimensions severely limit the use of radiated measurements in determing gasket performance.

The conducted energy method is described in SAE-ARP-1705: "Coaxial test procedure to measure the RF Shielding Characteristics of EMI Gasket Materials". This standard describes a test method for determining the "Transfer Impedance" of a gasket material using a small coaxial fixture and relatively simple instrumentation. In the transfer impedance test, illustrated in Figure 1, a signal is applied through a 50 ohm load to a gasket specimen clampled between 2 plates. Transfer impedance is proportional to the ratio between signal voltage and measured voltage across the test fixture. This method has excellent repeatability and is sensitive enough to detect small differences in performance between different materials, finishes, or loads.

The conventional coaxial test fixture described in SAE-ARP-1705 uses a pressurized diaphram to apply load to the gasket material. Load is calculated by multiplying diaphram area by a pressure gauge reading. Inaccuracies can result from diaphram stiffness, guage error, and poor regulator

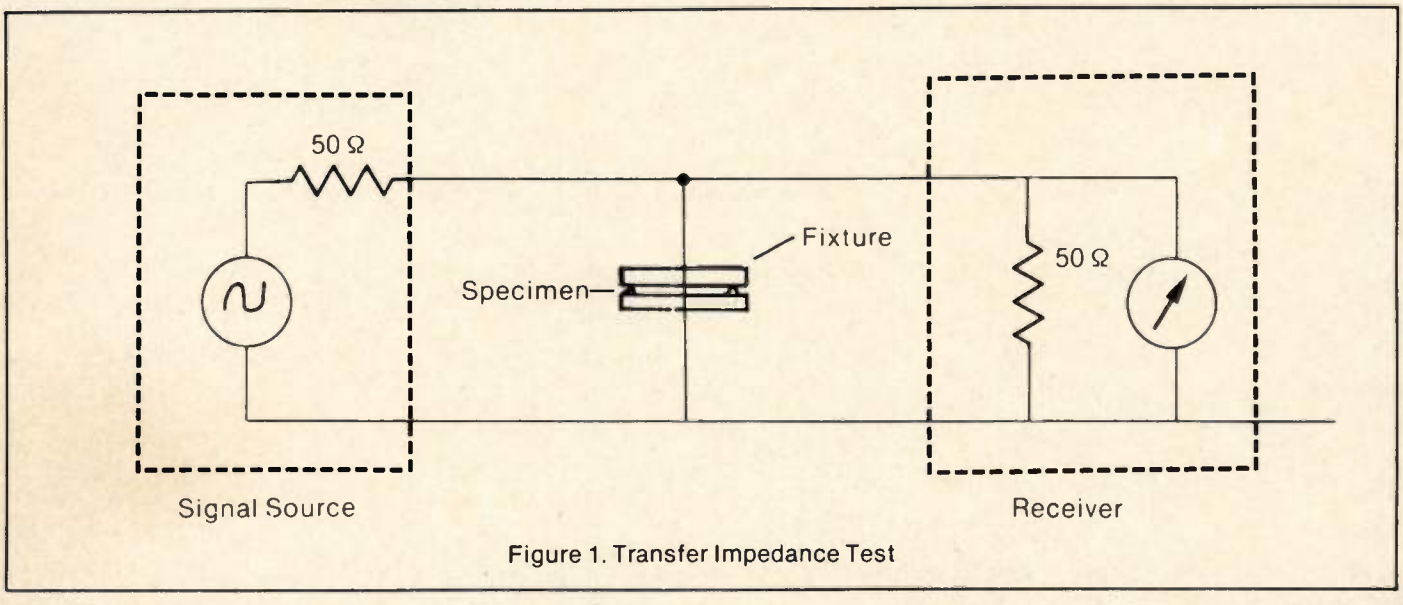
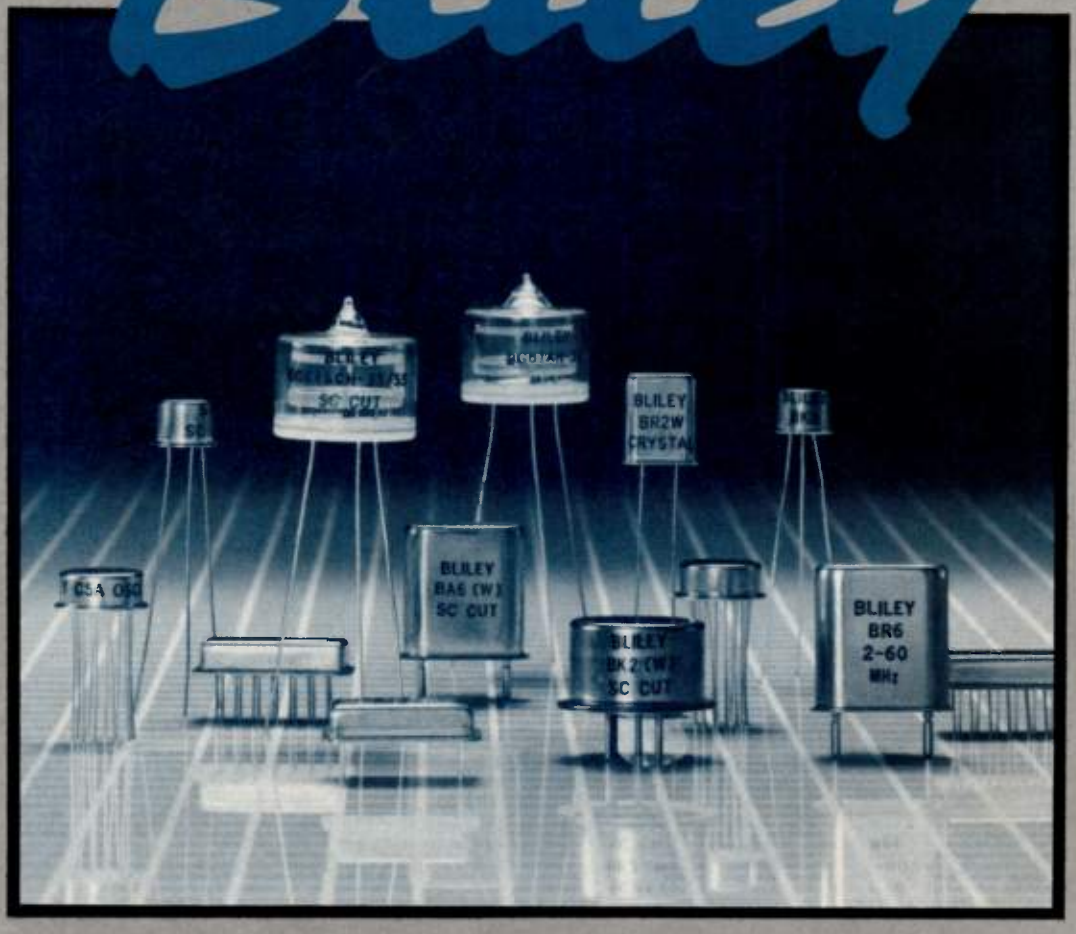


Figure 1. Transfer Impedance Test

Bliley



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performance. Tests on conductive elastomer gaskets requiring heavy loads produce an accumulated error that is insignificantly small. However, certain materials, such as Beryllium Copper finger stock, are used at compression loads typically one-tenth to one-onehundredth as heavy as conductive elastomers. In this light-load range, the accumulated error *is* significant, and cannot be compensated for by a calibration procedure.

Gasket loading of less than approximately 10 pounds per linear inch cannot be applied by the diaphram assembly because pressure regulators typcially will not work below a setting of 5 psi. The diaphram is also limited in usable defixtures for measuring high deflection, low force EMI gasten percent from free height, this limited range presents no problems. However, Beryllium Copper finger stock can typically be compressed 75% from free height, a distance which exceeds the diaphram range. Thus, expensive special tooling and repeated remounting of specimens are required to test these large deflection materials in the conventional fixtures.

Test Fixture Design

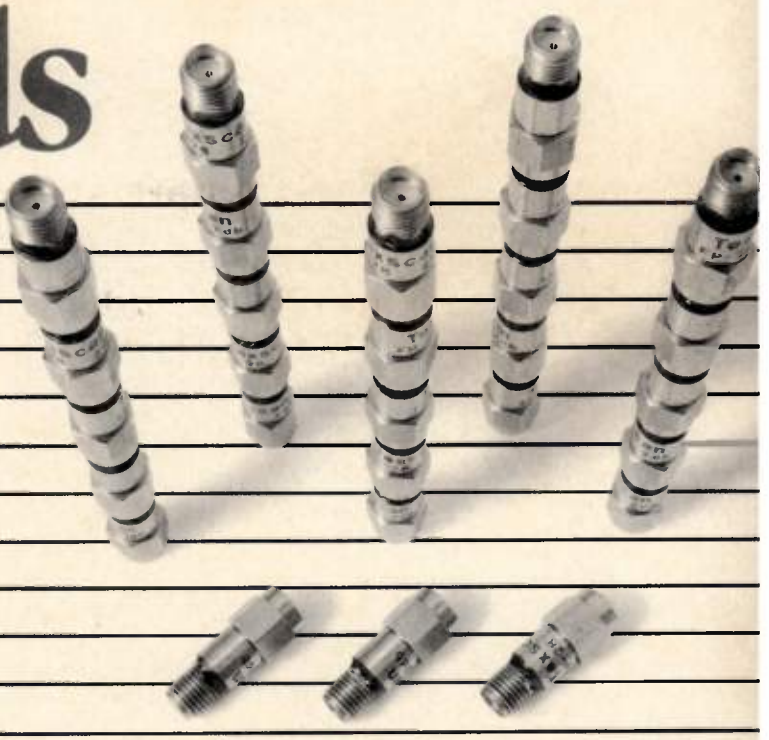
A new coaxial test fixture design was obviously necessary in order to make practical the testing of low force, large compression materials. Requirements included a specimen height range of 0-0.5 inches, a load range of 0-800 pounds, and the ability to calibrate load readings, particularly at light settings. A simple, inexpensive test plate design was considered desirable so that a variety of mating surfaces could be investigated. Such plates would accommodate a variety of gasket materials and allow quick changeover from one material to another.

Instrument Specialities Co., Inc. has designed and built a coaxial test fixture (Figure 2.) for transfer impedance measurement which meets the requirements for testing all types of shielding gasket materials including Beryllium Copper finger stock. The new fixture has the same 6" diameter as the standard fixture, for easy data correlation. Heavy wall welded steel construction was chosen to prevent mechanical distortion under heavy loads. Housing components are tin plated to assure corrosion free low impedance electrical contact. Three, 2 inch bore single acting pneumatic cylinders are used to apply load to the gasket material. Single acting air cylinders, unlike diaphrams, have return spring preload which must be overcome before any motion occurs. Preloading eliminates the problem of poor regulator performance below 5 psi, since the motion threshold occurs at a higher pressure. The three cylinders can apply both heavy enough loads to test conductive elastomers, and light enough loads to test Be-Cu finger stock within a height range of 0-.5 inches. The air cylinders are a reliable, inexpensive alternative to the troublesome diaphram assembly in the standard fixture.

Force is applied to the test plate at only three distinct points, and three load cells are used to measure the force. A load cell is mounted on each cylinder rod and can swivel to accommodate uneven test plate alignment. Unlike the diaphram assembly, the load cells provide direct force measurement. Calibration can be accomplished easily by placing an appropriate standard weight on the lower test plate and recording the load cell reading. If a gasket material is to be tested at a fixed height rather than a fixed load, non-conductive coaxial spacers are installed in the gap between the test plates along with the material specimen. Air pressure is set high enough to seat the spacer against the top plate, and the load reading is ignored.

Test plate design is greatly simplified in the Instrument Specialties Company fixture because no diaphram seal is needed. The standard lower plate is a disk with a threaded

Fixed Pads



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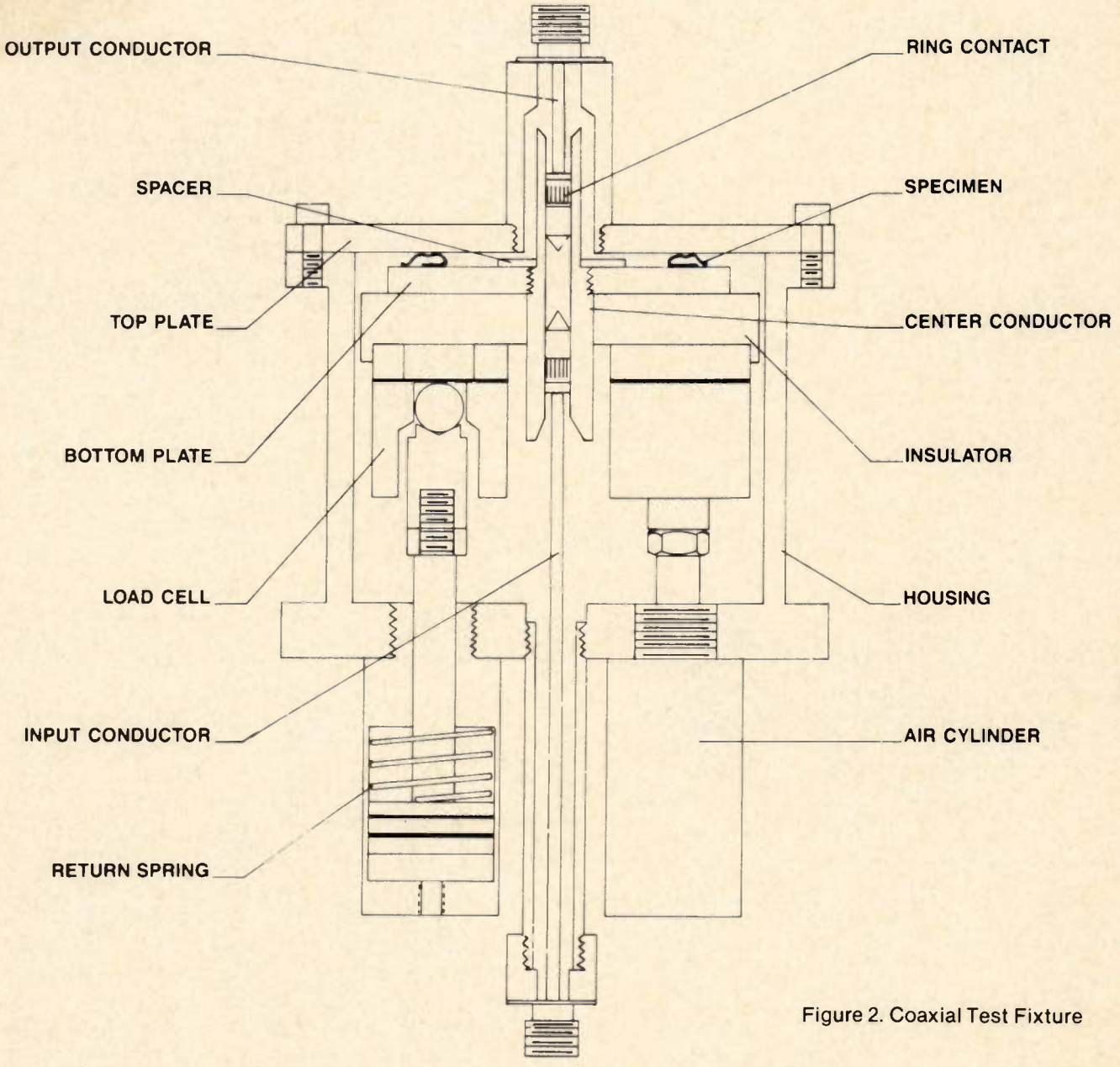

Figure 2. Coaxial Test Fixture

hole in the center. The upper plate is rectangular with a threaded hole in the center and 6 bolt holes around the perimeter. No other machining is necessary unless special fastener holes are needed to mount a specimen. Plate material can be any metal or conductive composite capable of withstanding the desired test load. Changeover time from one specimen to another is less than 5 minutes, excluding time to mount the specimen. RF continuity is maintained with the movable bottom plate by means of 2 silver plated Be-Cu finger rings sliding within the center bore of a silver plated center conductor. This conductor is screwed into the center hole of the lower test plate and is located by a phenolic insulation plate which is supported by the air cylinders. The fixture is closed by bolting down the top plate with 6 cap screws.

A variety of instrumentation packages can be used with the transfer impedance measurement fixtures, using either a manual or computer controlled system. A computer system requires less time to perform a test and provides a convenient vehicle for storage and retrieval of the resulting data. Since transfer impedance data is usually presented graphically, a computer-driven digital X-Y plotter can be used to good advantage. This test set up is shown in Figure 3.

All instrumentation systems for transfer impedance measurement include a signal source and some type of receiver. Both instruments should be capable of a frequency range of approximately 10 KHz to 1 GHz, since relative values can be measured over this range. Absolute accuracy, however, is limited by fixture resonance which first occurs at approximately 700 MHz. Signal generator requirements must include output amplitude at least 140 dB above the smallest signal which the receiver is capable of reading. For automated testing, the signal generator must have a con-

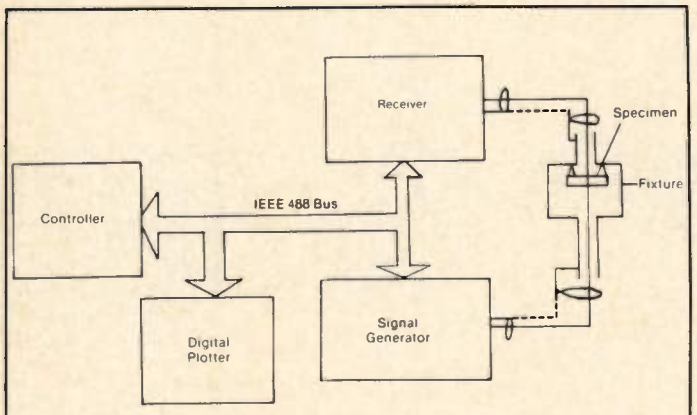

Figure 3. Test Set Up

stant voltage output, regardless of load. Suitable receiving instruments include spectrum analyzers, wave analyzers, and EMI receivers (tuneable RF microvoltmeter). If a computer system is to be used, IEEE 488 compatible instruments are most convenient. Interconnections are made with 50 ohm coaxial cable using type N connectors, and should be as short as possible to minimize signal losses.

Instrument Specialties Co., Inc. has run a series of Transfer Impedance tests using the new fixture design. Computer controlled instrumentation was used, with a spectrum analyzer as a receiver. Repeatability across the spectrum from 10 KHz to 1 GHz was found to be plus or minus 1 dB, which is approximately the accuracy of the instruments. With this level of repeatability, subtle differences in effectiveness between various specimen geometries, plated finishes, mating surface conditions, and test loads were observed. Tests were performed using beryllium copper finger stock, conductive elastomers, oriented wire mesh, and metal spiral gasket materials. Special tooling was required only for those materials which must be mounted with mechanical fasteners. Test duration, including setup, was only 30 minutes due to fast specimen changeover and automated instrument operation. Although most of the tests were run using spacers to control specimen height, some were run without spacers and loaded according to force readings. Load settings could be maintained anywhere between 2 and 100 pounds per linear inch. Tests were conducted over the full height range of 0 to 0.5 inches with no special tooling or setup procedures.

Conclusion

In actual use, the Instrument Specialties Co. transfer impedance test fixture has proven superior to conventional fixtures for measuring high deflection, low force EMI gasket materials. Since other materials can also be tested with equivalent accuracy and convenience the new fixture provides a nearly ideal method for evaluating the relative effectiveness of common EMI gasket materials. □

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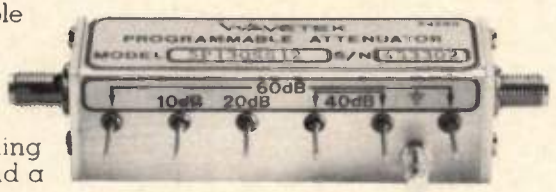

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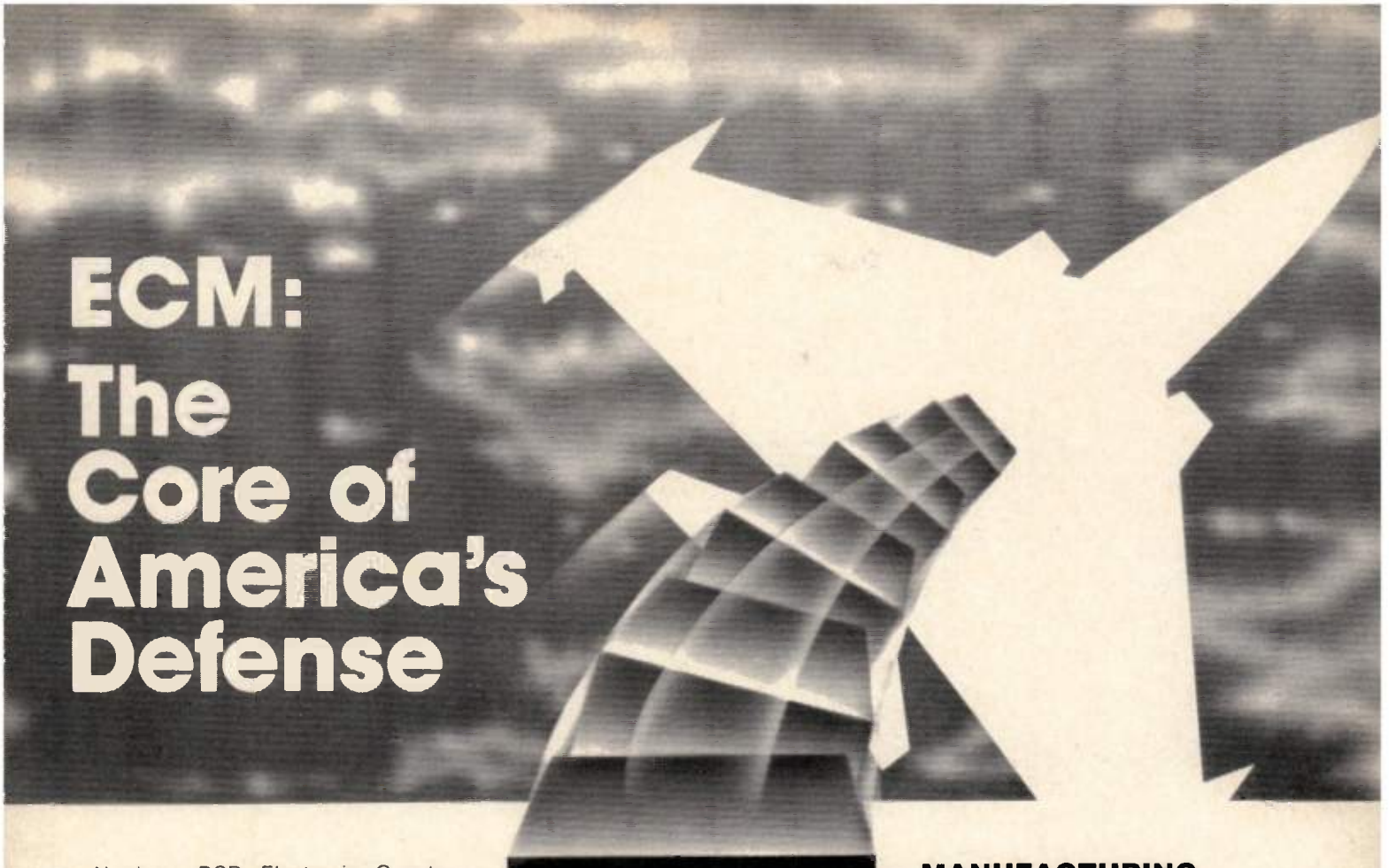

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High-performance network analysis for bench use or automatic testing is now simple and cost effective with Hewlett-Packard's new Model 3577A network analyzer. This analyzer expands testing capabilities for device characterization in the R & D lab and production test. High accuracy and resolution over the 5 Hz to 200 MHz frequency range are now available in a single product. Measurements can

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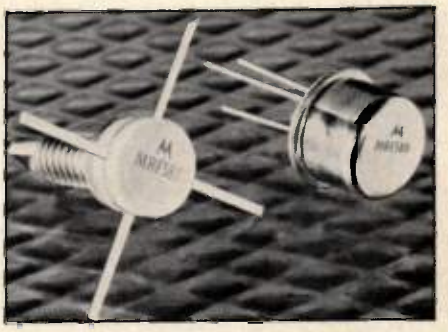
CATV RF Transistors

Motorola introduces two CATV RF transistors, the MRF586 and MRF587. These devices have guaranteed functional tests, such as noise figure and gain associated with noise figure. They are designed and characterized to be direct replacements for TRW's LT1001A and LT2001. The primary applications for the MRF586 and MRF587 are cable television distribution amplifiers and set top converters. They are also ideal for electronic instrumentation manufacturers who need wide band, low noise amplifiers up to 500 MHz. Pricing is:



MRF586 $2.20 (100-999), MRF587 $9.55 (100-999). Sample quantities are immediately available from warehouse stock, and production quantities will be available both from the factory and through authorized Motorola distributors in approximately 6 weeks from date of order. **Motorola Semiconductor Products Inc., Phoenix, AZ 85036, INFO/CARD #139.**

Linear Amplitude Modulator/Demodulator

The model 2A linear detector/modulator has been introduced by T-TECH Co. to tackle the common RF instru-

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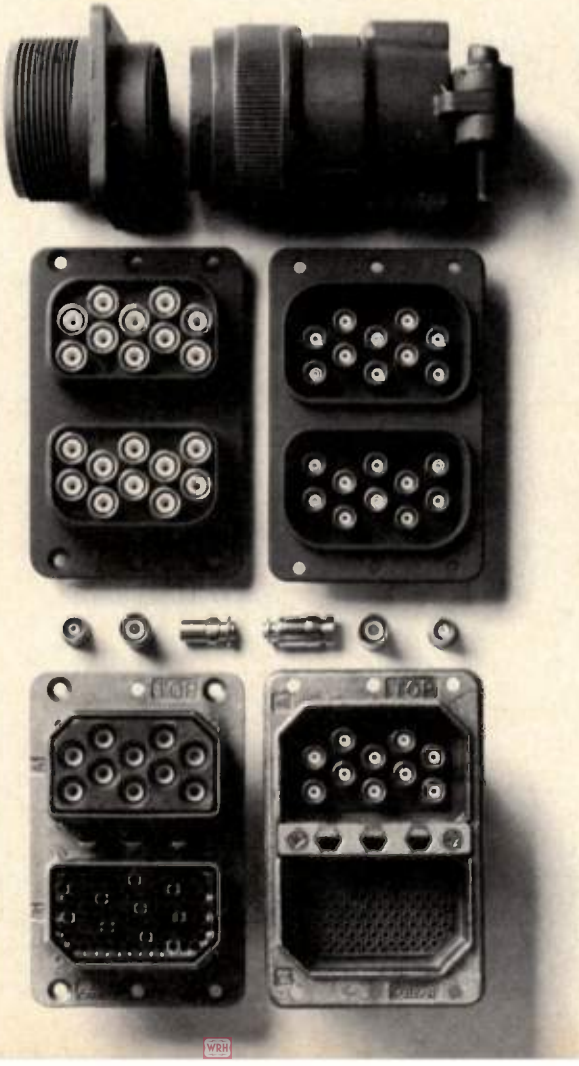

Our custom-designed PHP Series is a high-performance, low-insertion-force, blind-mate series of RF coaxial connectors which require no threading and are especially well-suited for countermeasure electronic equipment.

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mentation and communications problem of difficulty in obtaining linear AM detection and modulation in high-modulation index applications at high RF and baseband frequencies. The model 2A device uses a new patented technique of indirect baseband high-speed feedback. In one of its operational modes, the new T-TECH model 2A can be used as a "zero-distortion" amplitude modulator. Total harmonic AM distortion is less than 0.15% for a 95% modulated carrier using 1 kHz sinewave modulation. Modulation bandwidth is 200 kHz. Modulator output impedance is 50 ohms at up to 1 volt RMS, with at least 10% linear overrange. Input impedance is 1 meg ohm. In addition, the model 2A can be used closed-loop with its own special diodes, "zero-distortion" amplitude modulator, built-in servo loop and self-contained 10 MHz quartz carrier source to

form a "zero-distortion" amplitude demodulator or linear detector. Applications include a linear-scale RF voltmeter that can also demodulate the RF instantaneous envelope; AM shortwave broadcast modulation monitors; AM transmitter modulator distortion-cancellation feedback; signal generator linear levelling loop detector, making possible low-distortion AM at the RF output; precision AM demodulators that do not require input tuning anywhere to 1 GHz, thus eliminating asymmetric distortion caused by RF frequency-selective networks; swept-frequency test instruments; spectrum analyzers; radar AGC, STC, etc.; plus applications as a modulator-only for low-distortion requirements. The carrier frequencies that can be demodulated by the 2A in the demod mode range from 10 kHz to 1 GHz. The 2A is encased in a 4.95" long by 2" by 0.6 inch epoxy potted module with pins for printed circuit mounting. The unit price is $405.00 in moderate quantities. A modulator-only version will be available in the future, as will a 10 MHz to 18 GHz linear detector model. The 2A is totally self-contained for detection applications above a 10 MHz carrier frequency, except for a required external trimpot to null the servo loop offset. Circuitry on-board the module includes a 10 MHz quartz

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oscillator (with access via pins), internal frequency-compensation of servo loop for carrier frequencies above 10 MHz (lower carrier frequencies are accommodated by adding external compensating capacitances with a proportionate reduction in baseband bandwidth), a "zero-distortion" AM modulator, two matched specially-designed detector diodes (with access via pins) with internal broadband matching networks, and 50 ohm input termination. **T-TECH Co., Hudson, MA 01749, INFO/CARD #138.**

RF Power Amplifier

Trilectric Inc., announces a new RF power amplifier for repeater applications, designated the A-6100UR. The A-6100UR produces 70-100 watts output power when driven with 2-6 watts input. Operating in the 450-512 MHz

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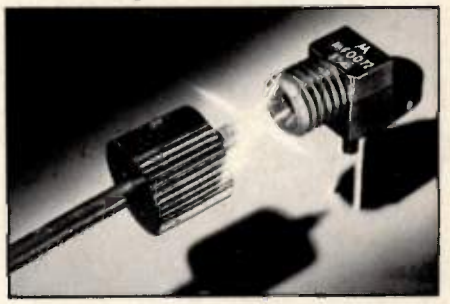
Low-Cost Fiber Optics System

A new and unique series of fiber optic components, designed to accelerate the utilization of fiber-optics technology in the high volume (industrial and automotive) marketplace,

has been introduced by Motorola. The series — a GaAs infrared emitter and a variety of four different detector configurations, when used with a 1000 micron core plastic cable, represents a matched system that offers significant advances to the current state of the art in each of three categories: Performance, Assembly Simplicity, and Price. Capitalizing on a recently installed and highly advanced GaAs facility, Motorola has developed a new emitter chip with an electrical bandwidth of 20 MHz at a wavelength of 820 nm. This wavelength represents an excellent match in the propagation characteristics of GaAs, silicon and plastic fiber, thus assuring maximum signal transmission. It makes the system useful over distances from 10 to 40 meters, depending on the type of detector employed. The detectors in-

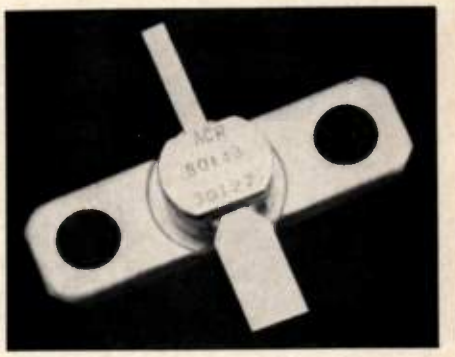


clude a very high gain photodarlington, a somewhat lower-priced phototransistor and a very fast PIN photodiode. Used in conjunction with the matched emitter, these detectors have a specified minimum response of 6 mA/μW, 100 μA/μW and 0.15 μA/μW respectively, at a distance of one meter. In addition, a Schmitt trigger photodetector will soon be available. Thus, the system can be tailored to a wide variety of applications through the choice of the appropriate detector. It is estimated that utilization of these components will result in a cost savings of approximately 75% of previously

available components with similar performance capabilities. Component prices are as follows: MFOE71, 820 nm Infrared Emitter $1.55, MFOD71, PIN Diode Detector $1.15, MFOD72, Transistor Detector $1.05, MFOD73, Darlington Detector $1.15. Prototype quantities of all devices except the Schmitt trigger are available at this time with full production quantities slated for availability in early 1984. **Motorola Semiconductor Products Inc., Phoenix, AZ 85036, INFO/CARD #135.**

High Gain Microwave Power Transistor

The 80143, a high gain, Class A, one-watt driver for general purpose broadband amplifiers, has been introduced by Acrian, Inc. The 80143



offers users three major advantages: high gain, high frequency cutoff and low cost. Acrian guarantees the power gain and output of the 80143 for 9 dB at 2.3 GHz. In addition to having high gain linear power, "the 80143 is designed for ruggedness, withstanding infinite VSWR loads." The 80143 is available immediately. In quantities of 500, the price is $27. **Acrian, Inc., Cupertino, CA 95014, INFO/CARD #134.**

Modular Measurement System

Because of the growing diversity and complexity of both analog and digital testing throughout the industry, Autek Systems Corporation has introduced a totally new measurement system that can be easily "customized" for a user to carry out digital, analog, RF and hybrid testing. Called the Series I Modular Measurement System, the system features plug-in modularity, which permits programmable interaction between digital, analog, coaxial and power test modules. Programming is accomplished through a specially designed 16-bit microcomputer controller which features a user-friendly higher-level language; graphics capabilities and a menu-driven test executive. A large and growing family of standard digital and analog modules allows the user to tailor the system configuration quickly and economically for a clean, quality "total" system. The new test head fixture for the Series I has been completely modularized as well, permitting test engineering to put the test drivers closer to the UUT rather than running long cabling which causes system response, crosstalk and noise

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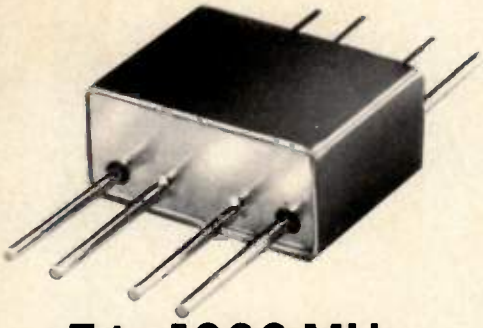

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	LO-IF	45	40
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problems. Utilizing this approach, the user can have up to 300 active pins at TTL speeds. For multi-logic applications the user can have 128 active pins in a single fixture configuration or 256 active pins in a double fixture configuration. **Autek Systems Corporation, Santa Clara, CA, 95051, INFO/CARD #133.**

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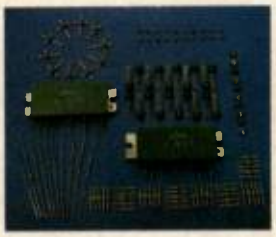

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Power transistors too! Power transistors up to 200w at 30 MHz and 40w at 470 MHz, are available, together with power modules for all mobile radio and amateur bands.

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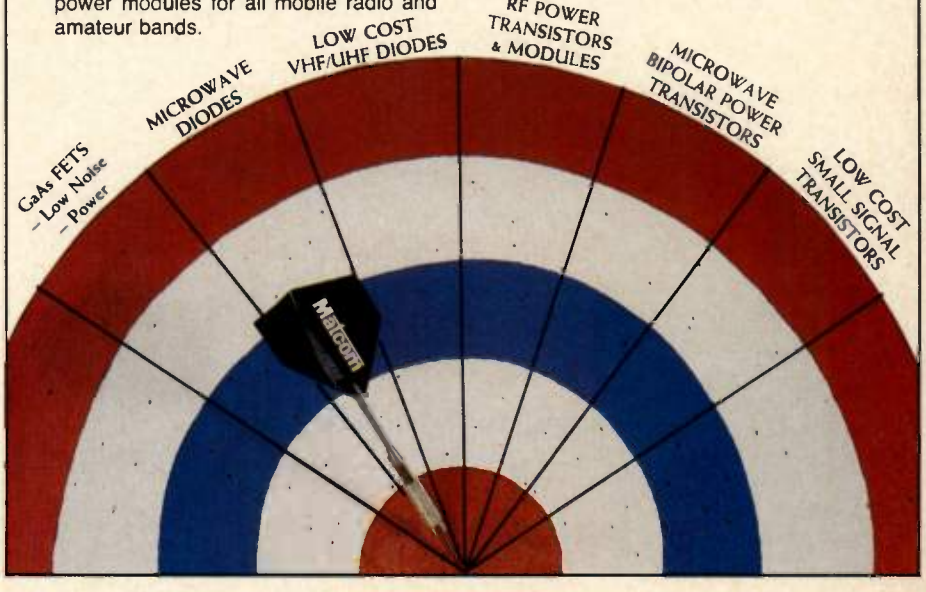

INFO/CARD 53

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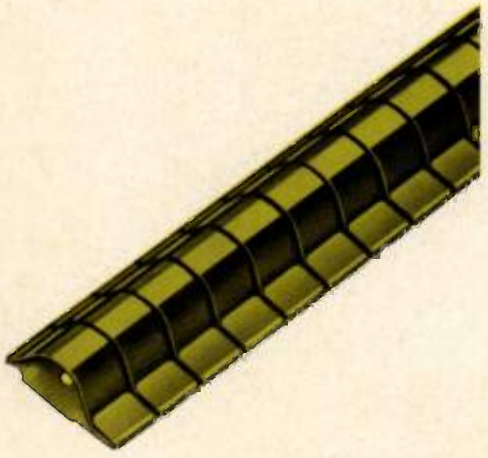


are available with screw sizes from #4 to 1/4" and hex sizes from 5/32" to 1/2". All are silver plated and include hardware. Proprietary resin formulation protects capacitor elements from the environment. **Spectrum Control, Inc., Erie, PA, INFO/CARD #131.**

Silicon Transistor Offers Low Cost Alternative to GaAs FETS

Silicon transistors with a line width of only 0.8 /um and an operating range of up to 6 GHz are available from Microwave Semiconductor Corporation. The devices offer a low cost alternative to gallium arsenide components within the 4 GHz range. The rated noise figure for the MSC BFQ 77 is 2.8 dB at 4 Ghz (8 dB associated gain). The CEREC package is suitable for mounting on all types of printed circuit boards. **Microwave Semiconductor Corporation, Somerset, NJ 08873, INFO/CARD #132.**

Active Filter Module

Sabor Corporation, Torrance, CA, has introduced a broad range, voltage tunable, active filter module designated VCF 40, will find wide usage in laboratories, testing applications, and OEM devices. Sabor reports the Model VCF (voltage controlled frequency) Filter Module is a low frequency active filter packaged in a small case. The center frequency of the filter can be tuned in direct proportion to an applied voltage. As the Q factor is adjustable, characteristics such as flatness, phase linearity and staggering can easily be adjusted by the user. With 4 external capacitors, the desired frequency range can be selected. The operating voltage is ±15 volts, and the module can be board mounted with other components. The VCF can be used as a low-pass, high-pass or band-pass filter, or a mix of these. Very fine adjustment of frequency is possible and the filter can be changed to provide up to 24 dB/octave slope. Cutoff frequency (or center frequency of band-pass) can be shifted 1000:1 by voltage tuning within a selected range. Maximum frequency is 50 kHz and signal frequency range is DC(LP) to 200 kHz (HP). **Sabor Corporation, Torrance, CA 90503, INFO/CARD #130.**

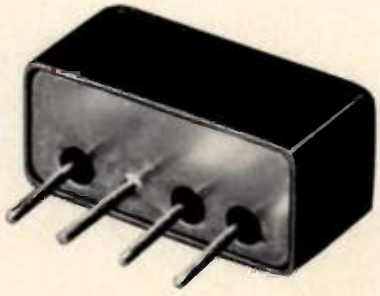
Automatic Dual-Sensor Power Meter

Combining two power sensors with the new HP 438A microprocessor-based power meter from Hewlett-Packard Company provides more than just two channels of RF and microwave power measurement. The user can compute the ratios A/B or B/A to display gain, gain compression or attenuation in dB or percent. Or, by using a dual coupler, the user can

measure and compute the power differences A-B or B-A to get the net power absorbed by a mismatched load. With a dual coupler and using the ratio mode, the HP 438A will indicate return loss in dB or power-reflection coefficient in percent. The HP 438A is optimized for ATE system applications, starting with its compact front

tiny power splitters

2 way 0°



1 to 400 MHz
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- **can be mounted upright or as a flatpack**
- low insertion loss, 0.8dB (typ.)
- hi isolation, 25dB (typ.)
- hermetically-sealed
- excellent phase/amplitude balance
- 1 year guarantee

TSC-2-1 SPECIFICATIONS

FREQUENCY (MHz)	1-400
INSERTION LOSS, dB (above 3 dB)	TYP.
1-10 MHz	0.25
10-200 MHz	0.4
200-400 MHz	0.8
ISOLATION, dB	25
AMPLITUDE UNBAL.	0.2
PHASE UNBAL.	2°
IMPEDANCE	50 ohms

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INFO/CARD 54 C93-3 REV. ORIG.

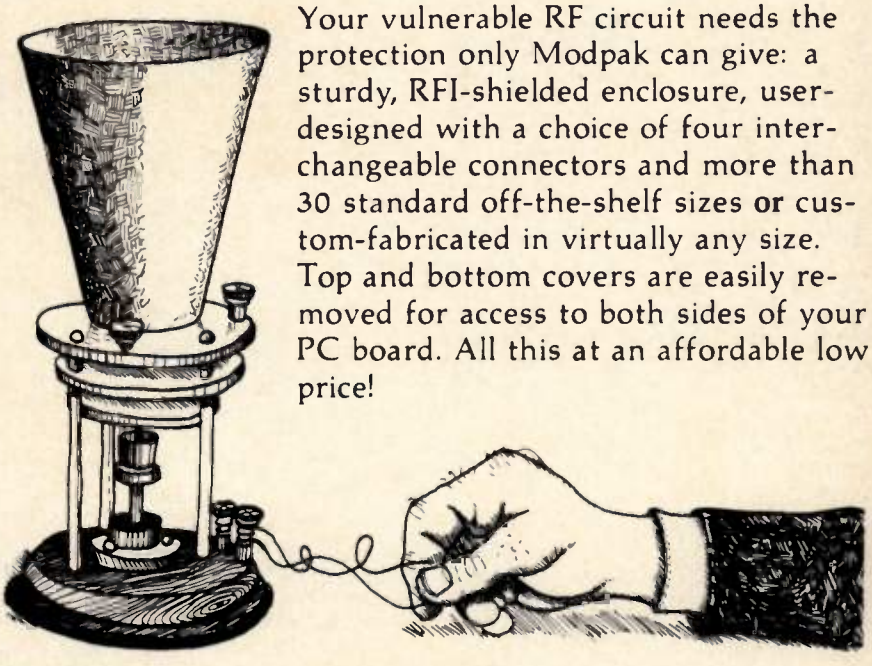
panel (89 x 213mm; 3.5 x 8.4 in.). For high-speed applications it can take up to 20 measurements per second. At slower measurement speeds, adjustable filtering provides up to 0.001-dB resolution over most of its power range, important in applications requiring measurement of small power changes. The HP 438A is $4,900. Option 002, rear panel connectors and oscillator, is $325. Sensor prices range from the HP 8482A (100 kHz to 4.2 GHz) at $580 to the high-power HP 8481B (to +44 dBm) at $1,402. **Hewlett-Packard Company, Palo Alto, CA 94303, INFO/CARD #129.**

3-1/2 Digit Micro-Ohmmeter

High accuracy measurements of low resistance values ranging from 0.00002 ohm (20 micro-ohms) to 200 ohms are easily made with resolutions down to 1 micro-ohm, by the new Ballantine 3205A Digital Micro-Ohmmeter. The Ballantine 3205A is a 3-1/2 digit instrument employing the proven 4-wire measurement method to cancel lead resistance errors inherent in less sophisticated 2-wire systems. Stability achieved in the 3205A circuitry allows quick, precise measurements to be made without the need

Bell didn't need Modpak™

But modern electronics involves more than a transmitter, a receiver, and a wire.



Your vulnerable RF circuit needs the protection only Modpak can give: a sturdy, RFI-shielded enclosure, user-designed with a choice of four interchangeable connectors and more than 30 standard off-the-shelf sizes **or** custom-fabricated in virtually any size. Top and bottom covers are easily removed for access to both sides of your PC board. All this at an affordable low price!

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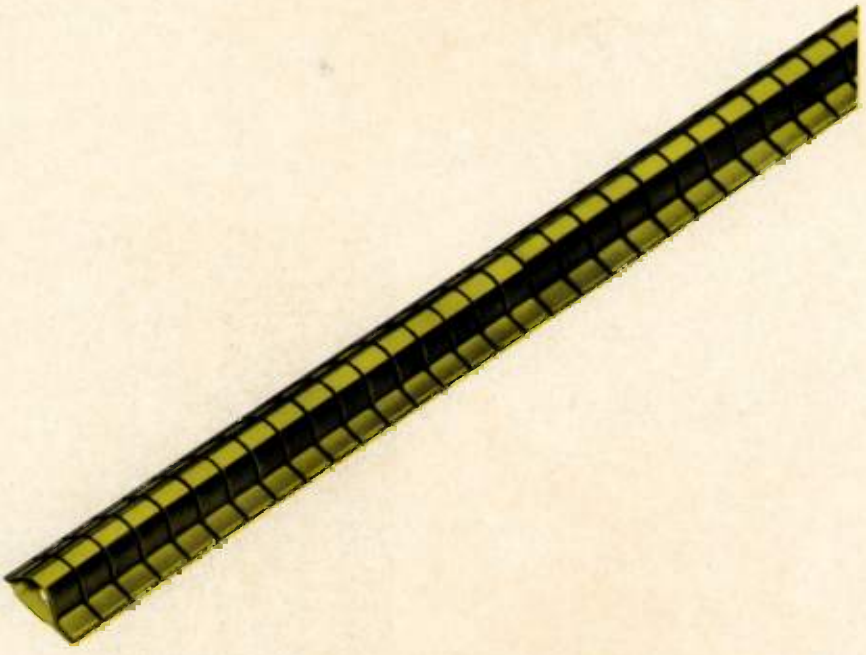

for time consuming and painstaking adjustments, commonly necessary when conventional resistance bridge techniques are used to measure low resistance values. The resistance under test is connected to the unit's constant current source by one pair of leads, and a second set is used to measure the voltage drop across the resistance inside the connection points of the current leads. This 4-wire

configuration eliminates the kinds of errors normally introduced in conventional 2-wire methods, where inherently the resistance of the test leads are included in the measurement. No corrections or adjustments to cancel extraneous lead resistance are necessary with the 3205A, the value read is precisely and only that of the resistance tested. For convenience in measurement applications, the 3205A is provided with a 4-wire set of Kelvin leads as a standard supplied accessory. The 3205A is in a multimeter-sized case, 8-3/4"W x 3-1/16"H x 10-3/4"D and weighs only 6 pounds. Price is $795, and the BCD Option 01 adds $125. Delivery is 30 days after receipt of order. **Ballantine Laboratories, Inc., Booton, NJ 07005, INFO/CARD #128.**

New Literature

Capacitor Catalog

A new 40-page Disc and High Voltage ceramic capacitor catalog is now available from Murata Erie North America, Inc. This new catalog, #58-05, provides complete technical information on the company's line of low voltage, temperature compensating, rectangular plate and medium to Hi K ceramic capacitors as well as safety recognized disc and high voltage ceramic capacitors. Also included are detailed specifications on taped and reeled disc capacitors for use with automatic insertion equipment. **Murata Erie North America, Inc., Marietta, GA 30067, INFO/CARD #127.**

RF Products Catalog

Watkins-Johnson Limited, located in Windsor, England, has released a new catalog covering its line of RF distribution products, electronic systems and antennas. Included in the catalog are descriptions and specifications of the company's antennas, dielectric masts, rotators, antenna systems, multicouplers, switching matrices, HF amplifiers and receiving systems. In addition, the 32-page document describes Watkins-Johnson Limited's capabilities in product design, repair and maintenance, post-design services and quality assurance. **Watkins-Johnson Company, Communications Department, Palo Alto, CA 94304, INFO/CARD #126.**

Distributor's Catalog

Microwave Distributors Company has produced a miniature handy reference catalog showing many of the products distributed by Microwave Distributors Company. The catalog stresses the quick delivery aspect of the company and sets forth what items are available from stock and the ability of the company to cross-reference competitor's products to those which are available off the shelf. The products stocked, in depth, by Microwave Distributors Company are manufactured by Solitron/Microwave, Applied Engineering Products (AEP), Amphenol, Kings Electronics, KDI

Pyrofilm, HH Smith, Insulated Wire, MIDISCO, and Aertech. **Microwave Distributors Company, Commack, NY 11725, INFO/CARD #125.**

Electron Tubes And Semiconductors Catalog

Richardson Electronics Ltd. offers its newest 24-page catalog listing its extremely broad line of over 12,000 different part numbers of electron tubes and power semiconductors. These parts are listed by type and price and are available on a guaranteed 24-hour order fulfillment basis, from Richardson's $15,000,000 inventory. **Richardson Electronics, Ltd., Franklin Park, IL 60131, please circle INFO/CARD #124.**

Electrical Mechanical Switches Catalog

M/A-COM, Inc. now has available its long-awaited electrical mechanical switch catalog. These switches, available with manual or remote operation, also offer a wide variety of connectors and actuation. M/A-COM's

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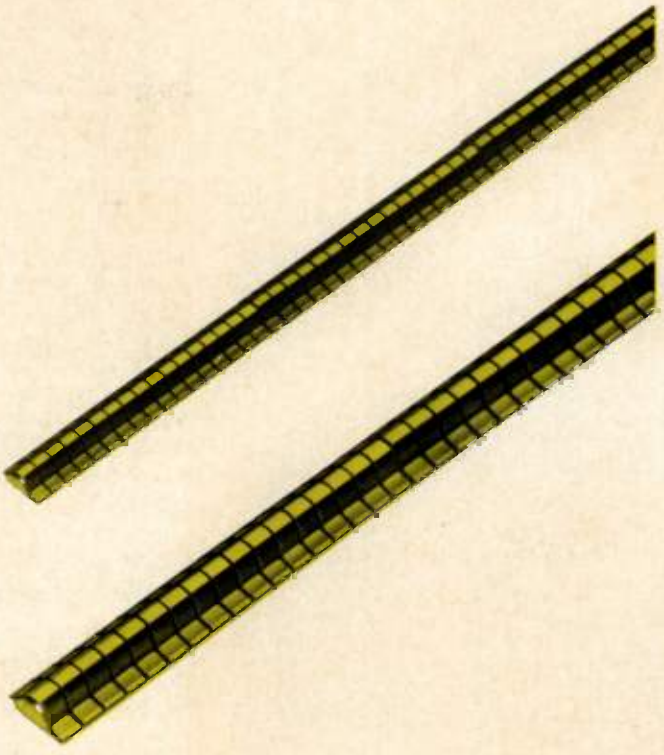


electrical mechanical switches are lightweight and compact, making them an excellent choice where these parameters are a design consideration. They also afford the user exceptionally fast switching times and are available with pressurization upon request. **M/A-COM Waveguide Products, Inc., Burlington, MA 01803, please circle INFO/CARD #123.**

Synergy Master Catalog

Synergy Microwave Corporation has just issued a Master Catalog covering its line of phase shifters, power dividers, mixers, modulators, transformers and multipliers. Included in this 44-page catalog is a section on Double-Balanced Mixers covering descriptions of: Dynamic Range, Harmonic Intermodulation, Two-Tone Intermulation, Conversion Loss and Noise Figure, Isolation, 1 dB Compression Point, 1 dB Desensitization Point and Return Loss/VSWR. A section is also devoted to Power Dividers/Combiners which includes descriptions of: Insertion Loss, Amplitude Balance, Phase Balance, Isolation, Internal Power Dissipation plus explanations of the device types available from In Phase N-Way Dividers to 180° Hybrids to 90° (Quadrature) Hybrids. The catalog gives features, guaranteed specifications, performance curves and outline specifications for each product line. **Synergy Microwave, Fairfield, NJ 07006, INFO/CARD #122.**

Communications Catalog

RF GAIN of Rockville Centre, NY, has just published the Flrst Edition of their new "Communications Catalog/ Cross Reference Guide". This first edition features expanded cross-references for Motorola, GE, Johnson, RCA/Tactec, Regency, Wilson, Quintron, Aerotron and NEW cross-references for Standard and Repco. Also included are the Japanese transistors for Midland, Force, Yeasu, Icom, etc. **Richardson Electronics, Ltd., Rockville Centre, NY 11570, INFO/CARD #121.**

Tools Catalog

The brand new Catalog 83-36S from OK Industries, Inc., features 145 pages of tools and equipment for electronics and telecommunications manufacturing, field service and labs, as well as schools and hobbyists. The large full-color, glossy catalog is divided into 9 sections covering a complete line of wire-wrapping tools, testing and troubleshooting tools, wire and cable, assembly products and aids of various types, and including N/C wire wrapping machines and support systems. The catalog also features a unique product line of low cost tools and products especially for educational and home use. Full of technical and product information, it is available upon request from **OK Industries, Inc., Bronx, NY 10475, INFO/CARD #120.**

Leader Instruments' Catalog

Leader Instruments Corporation has announced the availability of their completely new test and measurement instrument catalog, with twenty-five new product introductions. Released in August, the catalog contains eighty pages, and represents the most attractive product presentation in the most comprehensive format in the history of the company. **Leader Instruments Corporation, Hauppauge, NY 11788, INFO/CARD #118.**

Antenna Range Position And Control Systems Catalog

This 20-page catalog features a complete line of antenna positioning and control products manufactured by

Orbit Advanced Technologies, Ltd., Netanya, Israel. Included are single and multiple axis positioners up to 300,000 ft-lbs. **Flam & Russell, Inc., Horsham, PA, INFO/CARD #115.**

Crystal/Oscillator Catalog

Featuring AT-cut quartz crystals from 5 MHz to 150 MHz and crystal oscillators from virtually DC to 1 GHz for commercial and military applications. Oscillators are available in temperature compensated (TCXO), ovenized (OCXO), and voltage controlled (VCXO) designs. These products are for use in communications equipment with an emphasis on Microwave applications for crystals and high reliability military applications for oscillators. **Cinox Corporation, Cincinnati, OH 45232, please circle INFO/CARD #119.**

TaNFilm® Chip Carrier Resistor Networks Data Sheet

The Resistive Products Division of TRW offers product specifications for its recently introduced chip carrier resistor networks in a new data sheet. The new bulletin includes performance data, ordering information, test information as well as ratings and specifications. Each data sheet includes a photograph and outline drawings providing dimensional information. Schematic diagrams are also provided. **TRW Electronic Components Group, Corpus Christi, TX 78411, please circle INFO/CARD #111.**

High Reliability Power Supplies Catalog

An 8-page, 4-color catalog describing "Design-As-You-Order" miniaturized switching power supplies is now available from Arnold Magnetics Corporation. The unique Arnold System allows the user to select from pre-engineered AC or DC input modules with 1 to 10 DC output modules — up to 400 watts — to meet "custom" power supply requirements. The completed ultra reliable power supplies are furnished in encapsulated, miniaturized packages and meet MIL-STD-810C, MIL-E-5400 and MIL-E-16400. The new Catalog provides electrical and environmental specifications for more than 50 standard sub-modules. It includes a convenient "work sheet" that guides the user through the designing and ordering procedure. To simplify the process a completed example has

been worked out and presented in an "easy to follow" format. Also, a table of "Options" is provided and mechancial data is illustrated. Arnold Magnetics Corporation, Culver City, CA 90230, INFO/CARD #113.

Interference Control

A 16-page manual and catalog, "Guide to Interference Control Using Beryllium Cooper", has been issued by Instrument Specialties Company, Inc. This is the first time the problems of controlling electromagnetic interference have been compiled into one concise source. Sections include a technical discussion on design considerations and the technology involved in developing effective shielding, as well as presentation of formulas needed to calculate shielding effectiveness. In addition, the new manual illustrates and provides technical data on beryllium copper shielding strips, springs and rings for electromagnetic compatibility. Also included are important specifications on the new assemblies for grounding of high voltage static discharge with discrete points of contact. Instrument Specialties Co. Inc., Delaware Water Gap, PA 18327, please circle INFO/CARD #117.

Advertiser Index

New Books

Radio Communications Receivers

By Cornell Drentea

A Comprehensive Guide to Radio Receiver Design and *Technology* — This text is a sourcebook that covers the broad subject of radio receivers. It brings the history of receiver evolution, technical facts which have developed present radio technology, present and future trends in receiver design from system through packaging.

The book combines history, theory, application and construction details to make it a valuable reference on receiver design.

Available from TAB BOOKS Inc., Blue Ridge Summit, PA 17214, 288 pages, $19.95 hardbound, $13.95 paperback, 1982.

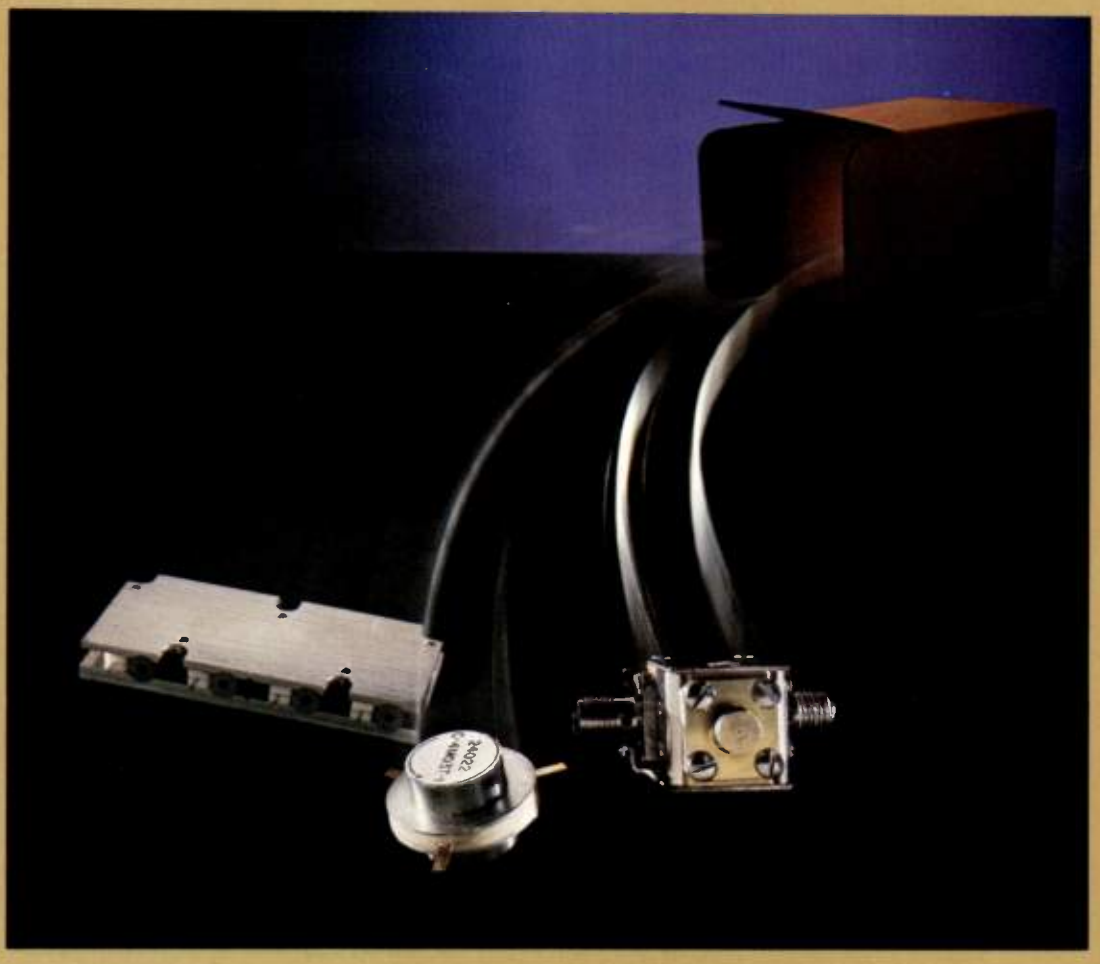
Transient Analysis Aided by Network Theorems

By Dr. Harry E. Stockman

This book is described by the author as *"The Engineer's approach to Transient Analysis"* or as *"A text dedicated to the time domain with its faithful servant, the D-Operator."* It contains simple ways of working transient problems in the time domain and s-domain, in the latter via Laplace's transform. Guidance is offered by means of "Data Panels" giving essential solution techniques by simple time saving procedures. A brief history with milestones of the development of operational and transform calculus is presented.

Available from Sercolab, P.O. Box 78, Arlington, MA 02174, 180 pages $12 plus $1.30 shipping, abroad add 10%.

Teledyne Microwave Delivers



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We take your needs seriously. With over 2000 documented isolator and circulator designs—including special mounting, multiport, high power, miniature package sizes, and every conceivable connector option—we're ready to customize a ferrite for your special needs. And we'll deliver on time.*

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Now is the time to rethink your design decisions— if you require up to 1 watt output for low-distortion intermodulation testing... broadband isolation... flat gain over wide bandwidth—or if you need much higher output level from your signal/sweep generator or frequency synthesizer — you can now specify Mini-Circuits' new ZHL-42 power amplifier... for only $895.

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ZHL-42 SPECIFICATIONS

Frequency	.7 to 4.2 GHz
Gain	30dB Min.
Gain Flatness	±1.0dB
Power Out @ 1dB CP	+29dBm Min.
VSWR In/Out	2.0:1 Max.
Noise Figure	7.5dB
Supply	+15V @ 690mA
Third Order Intercept	38 dBm Min.
Second Order Intercept	48 dBm Min.
Size	7" × 3¼" × 2⅛"h

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