THE RF CAPACITOR HANDBOOK

american technical ceramics

THE RF CAPACITOR HANDBOOK

THE RF CAPACITOR HANDBOOK

ATC 001-889 Rev. A
FIRST EDITION, REVISED

PUBLISHED BY:

AMERICAN TECHNICAL CERAMICS

american technical ceramics
one norden lane, huntington station, n.y. 11746-2142 usa
phone: 516-622-4700 • fax: 516-622-4748 • e-mail: sales@atceramics.com
http://www.atceramics.com

PREFACE

The purpose of this Handbook is to shed light on an area of technology in which there was formerly little or no information. In the early days of Radio when investigation into the characteristics of circuit components was being made, it was determined that the inductor was the main source of circuit losses, and that the contribution by the capacitor was negligible.

Although this was certainly true for narrow-band, vacuum tube voltage amplifiers up to a few Megahertz, today, high-power, low impedance, wide band transistors at many hundreds of Megahertz are commonplace, and this combination of microwave frequency, high RF current, and ultra-high current density due to ever smaller size has made the losses in the capacitor a significant factor in circuit efficiency. Even in cases where the inductor is still the dominating loss element, transistor gains are often so low (3 to 8. dB) that fractions of a dB in wasted RF power have become intolerable economically. In fact, at low impedance levels, such as the input to a power transistor, the current concentration in miniature capacitors so affects the capacitor's contribution to the loss of RF energy that it may exceed that of the inductor.

In the light of these problems, ATC undertook to provide the engineer with practical design data. To accomplish this:

Considerable research was done into present day literature on the effects of capacitors on RF circuit performance in an attempt to locate usable information. Little seemed available.

Ultimately it appeared that the uncondensed, oldest texts were often the most lucid in discussing circuit efficiency. Unfortunately, they neither treated capacitor types nor frequencies, impedance levels, or current densities appropriate to today's circuit techniques.

Therefore, Transmission Loss tests specifically oriented toward modern capacitors were formulated, fixtures designed and built, and tests run over a wide frequency range. The results of these tests gave rise to Application Notes which mathematically detailed the relationship between capacitor dissipation loss and the efficiency of tuned networks.

In contrast with a standard textbook presentation, this Handbook does not restrict itself in scope of content or audience technical level. It not only introduces the reader to the theoretical capacitor of a set of electrodes in a dielectric, but going further, examines the impact of basic capacitor characteristics on benchwork considerations important to the circuit designer.

As a result......

THIS HANDBOOK CONTAINS

- theoretical and empirical high-frequency circuit-design equations.
- characteristics of capacitor dielectrics, especially at high RF power levels and frequencies.
- actual test data at frequencies from 100 MHz to 3 GHz.
- practical design techniques for significantly increasing:
    - gain and power
    - DC-to-RF conversion efficiency
    - bandwidth
    - transistor lifetime
    - circuit MTBF
    - low-noise performance
    - your knowledge and understanding of high-frequency, high power effects on capacitors, and their influence on circuit design and performance.

We sincerely hope that this will provide you with practical, specific answers in many of the formerly "gray areas" of RF circuit design.

To reduce those areas of uncertainty which have cast the shadow of Alchemy across high frequency work, we offer to Handbook users the opportunity to contribute articles to the Second Edition which will tend to clarify UHF design for both the neophyte and the Senior Engineer.

ACKNOWLEDGEMENT

I would very much like to express my gratitude to the many microwave engineers who reviewed my original manuscripts for their critiques of the articles which now comprise the chapters of this Handbook.

I would also like to thank the magazine editors for their suggestions on arrangement and wording which helped provide a clearer exposition of the concept to be presented.

The research program of which this Handbook is an outgrowth required that high speed test fixtures be designed and built, and computer time purchased. Since we were exploring uncharted waters without software for the network analyzer that would simultaneously encompass all the desired tests, this program required considerable financial support. I am grateful that the management of ATC felt this pioneering effort was worth the sacrifice.

In the past when reading an author's acknowledgement of the year-long sacrifices his family had had to make during the writing of his book, I must confess I took it with a grain of salt. Now, having been through the experience, I certainly must say how grateful I am that my dear wife, Anne, lent her encouragement and patient support throughout this effort.

I hope you will find this Handbook as useful as we labored to make it.

Vincent F. Perna, Jr.

TABLE OF CONTENTS

Note: The dash-connected page numbers indicate Section, Chapter, and Page, respectively. Section 1 (Design Topic Locator) has been eliminated in this Fifth Printing.

TABLE OF CONTENTS

Note: Section 4 (ATC Capacitor Catalog and Product Data) has been eliminated from this Fifth Printing.

TABLE OF CONTENTS

SECTION 2

CAPACITOR INFLUENCE

ON

CIRCUIT DESIGN

RF CAPACITOR FUNDAMENTALS: A SYNOPSIS

The optimum design for capacitors used in tuning, matching, and coupling circuits operating at high power levels, or above about 100 MHz, requires a dielectric having extremely low loss at the frequency of use.

RF capacitor dielectric losses are described in terms of loss tangent (tan δ). Lower frequency capacitors are often characterized in terms of dissipation factor (DF). The two are actually equivalent, and different in nature from another low-frequency parameter: power factor. The higher the loss tangent, the greater the equivalent series resistance (R_S) of the capacitor to RF power, also: the poorer its quality factor (Q_{cap}), the greater its loss (heating), and the worse its noise characteristics.

A capacitor may de depicted by an equivalent resistance (R_S) in series with an ideal capacitance (C). R_S may be determined from: $R_S = (X_C/Q_{cap}) = (X_C)$ (tan δ). From this we can see that capacitors with lossy dielectrics and small values of capacitance will be highly resistive to RF energy.

The fact that tan δ and DF typically increase with rising frequency (and temperature), and that capacitance values employed commonly decrease with increasing frequency (not only for tuning but also to avoid resonance problems contributed by their own internal inductance), accounts for the serious capacitor heating encountered in RF amplifier designs employing low Q capacitors.

If the power input is great enough, the capacitor may fail due to internal thermal run-away, often taking the transistor with it. At best, if the power level is low, low Q_{cap} results in degraded stage gain and instability with temperature, requiring extra stages to make up for it. This is especially true where circuit resistance levels are low (e.g. the transistor input) and where the current has increased to the level of Amperes from matching down from Z_o=50.Ω. For example, at 300. MHz a 30. pF capacitor with low-frequency dielectric (e.g. K1200 material) will exhibit nearly 1. Ω of RF resistance. Whether used for series or shunt matching, extremely poor stage performance will result, since the base-emitter diode may also exhibit the same level of resistance. Dissipation Losses of from 0.5 to 3.0 dB per capacitor are not uncommon in lossy dielectrics at high power levels. The losses affect both gain and power, and can show up equally seriously as degraded Noise Figures in low-noise front-ends. Unfortunately, the engineer usually finds these facts out after he begins trouble-shooting his assembled breadboard. Although such a procedure is wasteful and frustrating, he generally does not have the time, budget, and special equipment necessary to investigate every potential circuit component before installation.

As an aid to the design engineer, ATC has developed a wideband test fixture to measure the (series-element) VSWR, Insertion Loss and S-Parameters of a range of capacitance values likely to be highly useful to the engineer. The chips (ATC 100, case A) were mounted on 25 mil microstrip lines on 25 mil alumina, and the data obtained on a Hewlett-Packard 8542A Network Analyzer (computer controlled; synthesizer referenced; phased locked; line-printer output).

The test data in Section 3 of this Handbook is a computer print-out of the Insertion Loss, VSWR, and S-Parameters to 3. GHz for a variety of capacitance values.

A Guide to Judging Microwave Capacitors

Here's how to evaluate the performance of the device before you put it into the circuit, by considering loss tangent, dielectric, and insertion loss.

PICKING the right capacitor for a given design becomes trickier with every advance in materials technology. Ideally, the capacitor should be small. Yet, if it heats too much, it will lose power and gain. Loss that normally occurs as heat can be converted to power and gain in the amplifier if the capacitor dielectric is chosen for low dissipation factor.

Advances in materials technology have produced a wide choice of capacitors in a variety of sizes with high dielectric constant. But which device is best?

A study of microwave capacitors profitably begins with an examination of the published loss tangents of different dielectric materials. At a single frequency one loss tangent may be 0.0001 and another 0.0002—a fact that is not very interesting until one discovers that the capacitor with the larger loss tangent would get twice as hot as the other.

Heat producers unmasked

The tangent value is only an indicator of relative heat loss. It points to the amount of series resistance that causes heating.

At frequencies below self-resonance, a capacitor is considered to consist of a pure capacitance, C_s, in series with pure resistance R_s, and both of these are shunted by a parallel pure resistance, R_p. The series resistance is the more important of the two because it carries most of the current. Typically low in value, R_s is between 0.01 and 2.0 ohms. R_p consists of the combined leakage paths and is of the order of millions of megohms. At impedance levels common to transistor circuits, R_p is ignored; R_s causes losses.

Losses of the capacitor occur primarily in the dielectric, the medium for energy storage and transfer. The quality factor of the dielectric, Q, is defined as the ratio of the amount of energy stored to the amount dissipated per cycle.

Dielectric materials are made of atomic charge carriers which are displaced from their original position when influenced by an electric field. Many dielectric materials contain dipole molecules that are naturally polarized. The dipoles, when influenced by an external electrical field, rotate until they are aligned in the direction of the field.

When the electric field alternates, the dipole molecules rotate at a rate proportional to their frequency of alternation. The limited molecular motion converts electrical energy to heat energy. The rate of conversion is generally in proportion to the power and frequency of the applied microwave energy. However, when the energy enters a dielectric it is attenuated at a rate proportional to the frequency of the electric field and the loss tangent of the material. Thus, if a capacitor stores 1,000 joules and dissipates only 2 joules in the process, it has a Q of 500. R_s is responsible for the energy dissipated and is directly proportional to the dissipation factor, DF. But how can the dissipation factor be made nearly nonexistent?

The ideal capacitor

The forward transfer coefficient of a capacitor used as a series element in a signal path is measured as a function of the dielectric phase angle, Θ. This angle is a measure of the difference in phase between the sinusoidal voltage applied to the capacitor plates and its current component. In low-loss capacitors, Θ must be very close to 90°. In the ideal capacitor, Θ equals 90°.

The dielectric loss angle, δ, is the difference between Θ and 90°. In microwave measurement of capacitors, the largeness of Θ is not as significant as the smallness of δ. A convenient measure of δ is $\tan \delta$, which goes to zero as the losses go to zero; hence the name "loss tangent."

The measure of δ is complex and is related to temperature and frequency.

Loss is a multi-function variable

The dissipation factor of a dielectric is also the tangent of the dielectric loss angle and is represented by the symbol DF. These terms are used interchangeably in the literature.

At a fixed temperature, $\tan \delta$ of a material varies with frequency and can be plotted on an

Vincent F. Perna, Jr. American Technical Ceramics, Huntington Station, N.Y.

Reprinted from MicroWaves, August 1970

X − Y graph. R_s is a function of frequency, capacitance and tan δ. Tan δ varies nonlinearly with frequency. A capacitor with a value of R_s when inserted into a circuit, will exhibit a proportional decrease in output power. This decrease is called "insertion loss."

Common dielectric materials used in capacitors have a high insertion loss at microwave frequencies. Instability occurs because the loss tangent rises with temperature and frequency. This frequency-dependent loss can result in thermal runaway, causing destruction of the capacitor and other circuit components as well. At low power levels, stage gain is degraded by "lossy" capacitors.

Internal heating due to power level

Loss tangent increases with temperature. Capacitor temperature is the sum of the ambient and any contribution made by internal heating due to losses. The magnitude of the internal contribution is a function of the rf power level to which the capacitor is subjected. Vacuum-tube circuits that operate at relatively low frequencies require capacitors capable of handling large voltages and moderate currents. The dissipation of these capacitors is expressed in the following equation:

$$W = E^2 \epsilon_r \tan \delta f \ (55.5 \times 10^{-6}),$$

where

W = watts dissipated per cubic centimeter
E = rf potential in volts rms
ϵ_r = the dielectric constant
f = frequency in Hertz
tan δ = loss tangent.

The reduction of any of these factors will reduce the dissipated power.

Capacitors in microwave transistor circuits, in contrast with vacuum-tube circuits, must handle large currents. Therefore the power loss must be determined in terms of current. The internal heating of the capacitor differs depending on whether it is made of low-frequency or microwave dielectrics (Table 1). Due to the extremely large impedance transformation, current in the base of a microwave transistor may be in the order of amperes.

The relative power dissipation of low-frequency and microwave dielectric materials is directly proportional to the loss tangent or dissipation factor. Tables published for various dielectric materials do not carry data much above 100 MHz because losses rise rapidly above this frequency. Table 2 shows losses for dielectric materials suitable for use up to 1 GHz.

Many capacitors suitable for use at moderately high frequencies, but not specifically designed for microwave applications, cause large insertion loss at ultra-high frequencies. This simply translates to power and gain loss per stage, resulting in poor circuit efficiency. Capacitor characteristics can be measured rapidly and accurately by means of S parameters. These figures provide the designer with a choice of capacitors at microwave frequencies.

Two key S parameters

Any two-part network may be described in terms of four parameters: input reflection coefficient, S_{11}; forward transfer coefficient, S_{21}; reverse transfer coefficient, S_{12}, and output reflection coefficient, S_{22}.

Low-loss capacitors are described adequately by two of these coefficients: reflection and forward transfer.

Table 1. Power loss at 300 MHz is compared

Parameter	Low Frequency Dielectric	Microwave Dielectric
DF (tan δ)	0.04	0.0001
$Q = \frac{1}{DF}$	25	1000
C	30 pF	30 pF
$X_c = \frac{1}{2\pi fC}$	17.6 ohms	17.6 ohms
$R_s = \frac{X_c}{Q} = X_c DF$	0.7 ohm	0.0176 ohm
Power dissipation @ 1.2 Amps	1.008 W	0.0253 W

Table 2. Loss characteristic of different dielectrics at 100. MHz

Dielectric type		tan δ = DF
K8000*	Barium titanate	0.1
K1200	Barium titanate	0.03
K30	Ceramic	0.002
K9	Alumina	0.0005
K15	Porcelain	0.00007

* K is dielectric constant

The ratio of reflected and incident power, S_{11}, for a bilateral passive network is the same as the voltage reflection, Γ, used in the calculation of VSWR.

The forward transfer coefficient, S_{21}, used as a voltage-ratio measure of transmission, has an absolute magnitude of slightly less than unity for capacitors of good grade. A value of S_{21} less than unity indicates unsuitability for use in low-noise or power applications. A figure of 0.943 is equivalent to an insertion loss of over 0.5 dB.

The forward transducer power gain, G_t, is obtained from $(S_{21})^2$ which, in the measurement of the passive network, may be converted to an equivalent power loss:

$$G_t = 10 \log_{10} |S_{21}|^2.$$

The size and physical construction of a capacitor may so disturb the electric field distribution on a transmission line as to cause an impedance bump or mismatch at that point. This, and the series resistance internal to the capacitor, changes the impedance of the line at the input to the capacitor. This results in a reflection of power toward the generator. The sum of deleterious effects is the insertion loss and is usually expressed in dB.

Analyzer measures S parameters

Computer-operated analyzers characterize two-port networks at microwave frequencies with

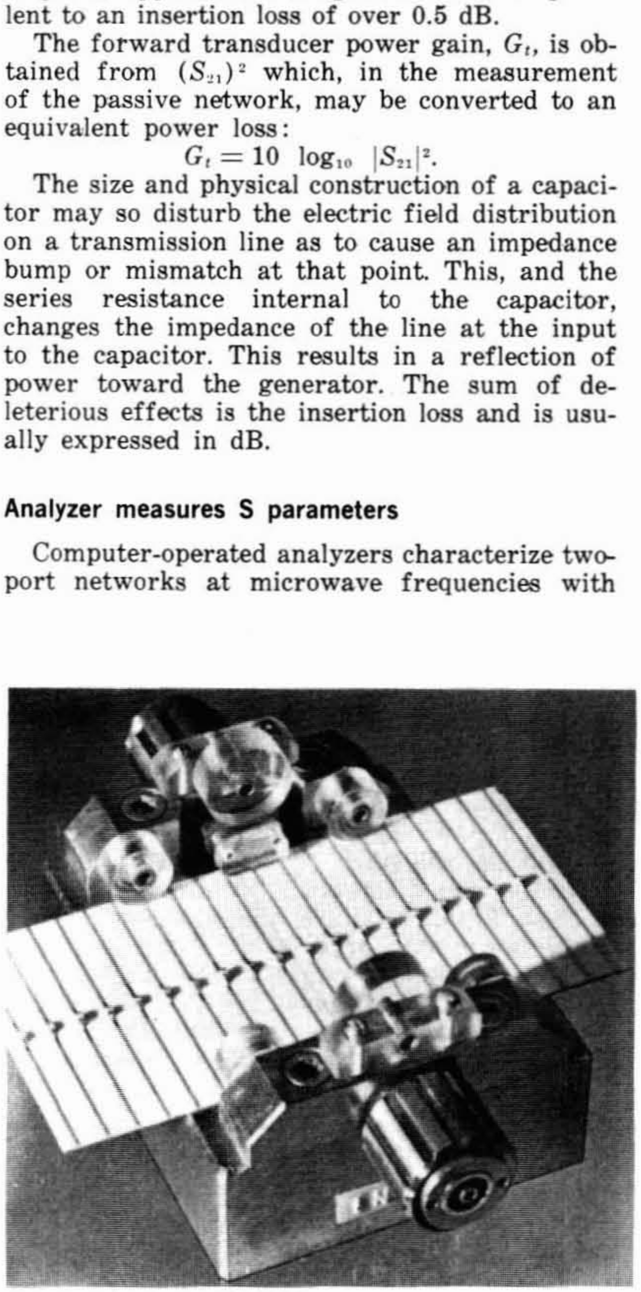


Fig. 1. Test jig speeds parameter measurement of capacitors by uniform feed. Cam-applied pressure assures low contact resistance.

Table 3. Computer readout prints S-parameter values

Freq. MHz	VSWR	Gain, dB	S_{11}	α_1*	S_{21}	α_2**
300.	1.116	−.01	.055	−81	.999	2
350.	1.100	.02	.048	−78	1.002	1
400.	1.090	−.01	.043	−75	.998	1
450.	1.082	−.02	.039	−72	.998	0
500.	1.077	.02	.037	−69	1.002	0
550.	1.072	−.01	.035	−65	.999	0
600.	1.068	−.02	.033	−63	.998	0
650.	1.064	−.02	.031	−60	.998	0
700.	1.060	−.01	.029	−58	.999	0
750.	1.057	−.02	.028	−56	.998	0
800.	1.054	−.02	.026	−55	.997	0
850.	1.050	−.02	.024	−55	.998	0
900.	1.047	−.03	.023	−56	.996	−1
950.	1.043	−.02	.021	−58	.997	−1
1000.	1.039	−.01	.019	−61	.999	−1

* Phase angle of reflected wave, α_1 varies with resistance, reactance and structure.
** Phase angle of transmitted wave, α_2 is influenced by path length, resistance and reactance.

such accuracy that stable circuits can be designed with ease. Data that accurately describes component characteristics in parameters suitable for computer readout are especially useful to the designer.

A typical computer readout (Table 3) has the capacitor characteristics tabled according to the following headings:

Frequency, VSWR, Gain, S_{11} and S_{21}, in which frequency is stepped in 50-MHz increments from 100 to 1000 MHz or any other range of interest. The S-parameter values are printed in magnitude and angle.

A test jig loaded with capacitors is shown in Fig. 1. Each table of measurements for a capacitor is printed in less than 60 seconds. The characteristics of the test fixture are measured and used as a zero reference for all subsequent measurements. Eccentric cams press the launcher center, conductors to the microstrip lines, maintaining low contact resistance.

Bibliography

Hausner, H. H., "Dielectric Calculations", *Ceramics Age* (April, 1947), pp. 165-166.

"S-Parameter-Circuit Analysis and Design," Hewlett-Packard Application Note 95.

Bauhaus, R. H. and Wong, R. W., "Predicting Transistor Behavior at the Threshold of Large Signal Conditions." *MicroWaves* (October, 1969), p. 40.

Froehner, W. H., "Quick Amplifier Design with Scattering Parameters," *Electronics* (October 16, 1967), pp. 100-109.

"*Ceramic Data Handbook*, Canners Publications (1963-1964), pp. 403, 405.

West-Cap Ceramic Chip Capacitor Handbook H31570, San Fernando Electric Mfg. Co., San Fernando, Calif.

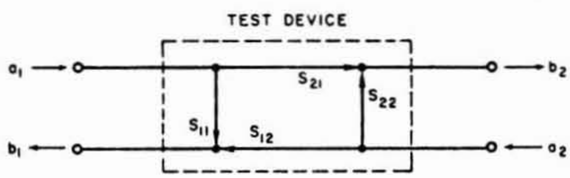
Valley, Jr., G. E. and Wallman, H., "Vacuum Tube Amplifiers," *M. I. T. Radiation Laboratory Series*, XVIII, p. 173.

What are s parameters?

S parameters[1,2] are reflection and transmission coefficients. Transmission coefficients are commonly called gain or attenuation; reflection coefficients are directly related to VSWR and impedance.

Conceptually, s parameters are like h, y, or z parameters insofar as they describe the inputs and outputs of a black box. But the inputs and outputs for s parameters are expressed in terms of power, and for h, y, and z parameters as voltages and currents. Also, s parameters are measured with all circuits terminated in an actual characteristic line impedance of the system, doing away with the open- and short-circuit measurements specified for h, y or z parameters.

The figure below, which uses the convention that a is a signal into a port and b a signal out of a port, explains s parameters.



In this figure, a and b are the square roots of power; $(a_1)^2$ is the power incident at port 1, and $(b_2)^2$ is the power leaving port 2. The fraction of a_1 that is reflected at port 1 is s_{11}, and the transmitted part is s_{21}. Similarly, the fraction of a_2 that is reflected at port 2 is s_{22}, and s_{12} is transmitted in the reverse direction.

The signal b_1 leaving port 1 is the sum of the fraction of a_1 that was reflected at port 1 and what was transmitted from port 2.

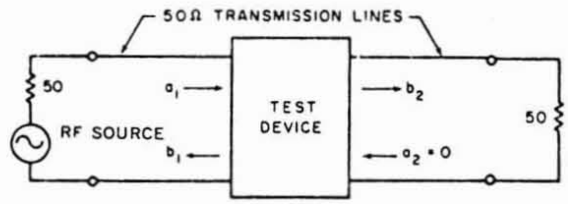
The outputs related to the inputs are

$$b_1 = s_{11}\, a_1 + s_{12}\, a_2, \quad (1)$$

$$b_2 = s_{21}\, a_1 + s_{22}\, a_2. \quad (2)$$

When port 1 is driven by an RF source, a_2 is made zero by terminating the 50-Ω transmission line, coming out of port 2, in its characteristic impedance.

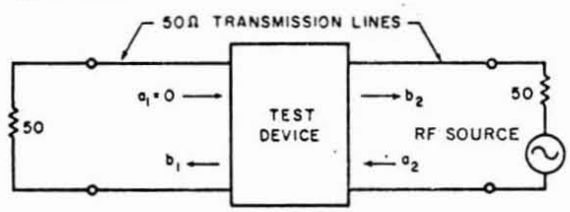
The setup for measuring s_{11} and s_{21} is this:



If $a_2 = 0$, then:

$$s_{11} = b_1/a_1,\ s_{21} = b_2/a_1 \quad (3)$$

Similarly, the setup for measuring s_{12} and s_{22} is this:



If $a_1 = 0$, then:

$$s_{12} = b_1/a_2,\ s_{22} = b_2/a_2 \quad (4)$$

Another advantage of s parameters is that, being vector quantities, they contain both magnitude and phase information.

By definition, s_{11} and s_{22} are ratios of the reflected and incident powers, or exactly the same as the reflection coefficient, Γ, commonly used with the Smith chart. The input and output parameters of a two-port device can be presented on a polar display without any transformation (see photos in text) and the corresponding normalized impedances can be readily obtained on the same chart. Impedance transformation and matching can be done either graphically or analytically. Mismatch losses that occur between any port and a 50-Ω termination can be calculated. For example,

$$P_{\text{Mismatch}} = 10 \log_{10} (1 - |\Gamma|^2),$$

where P_M is the mismatch loss in dB at any given port having a reflection coefficient Γ. When the s parameters are known, s_{11} or s_{22} can be substituted for Γ.

The transducer power gain of the two-port network can be computed by

$$G_T = |s_{21}|^2 \quad (5)$$

or in dB

$$G_T = 10 \log_{10} (|s_{21}|^2) \quad (6)$$

Reprinted from "Combine S-parameters with Timesharing" by Les Besser, Electronic Design 16, August 1, 1968, pages 62-68.

S-PARAMETERS:
PRACTICAL ASPECTS FOR BOTH RF AND DIGITAL

by

Vincent F. Perna
Vice President
Microwave Engineering
American Technical Ceramics

(An extract from the paper "High Bit-Rate Hybrid Circuits and Components as Seen from the Frequency Domain", delivered at the November, 1970 ISHM Hybrid Microelectronics International Symposium in Los Angeles)

The rapid advance of digital technology into the region of picosecond rise-times and Gigahertz data rates has eroded the formerly clear-cut distinction between digital and microwave circuit designers. Some disciplines, such as computer sorting of detected radar pulses, for example, are *completely* indistinguishable.

Digital engineers considering rise times of the order of 50 ps. must be well versed in VSWR, reflection coefficient, and transmission-line impedance---once considered the exclusive domain of the microwave engineer. For rise-times equivalent to several GHz of bandwidth, he must cope with such long-time microwave problems as component package parasitics, impedance transformation, unexpected coupling, and the uncertainties of microstrip transmission coefficients.

Fortunately, microwave component manufacturers now offer the digital designer rational and convenient methods of component characterization which account for an otherwise confusing array of invisible and stray influences viewed from the time domain.

Capacitors, for example, above about 1 GHz (that is, rise times $\cong \frac{0.35}{BW} \cong 350$ ps.), are no longer looked upon as a simple assemblage of plates, but are described in terms of their scattering-parameters in series or shunt on a transmission line.

Consequently, to take advantage of these useful characterization methods, the rise-time and pulse-width engineer who would work with the input, output, or just internal transfer of data at multi-megabits per second must become a practitioner of the semi-heuristic art of microwaves, where problems such as terminating a transmission line over a wide bandwidth without increasing power supply drain have already been solved.

Over a certain bandwidth, this may be accomplished by employing physically minute, but high-valued, capacitors with extremely low inductance.

To determine what this usable range is, requires an examination of a proposed capacitor's transmission coefficient and reflection coefficient versus frequency. "S-Parameters" are used, since they are far easier to measure than H-, Y-, or Z-Parameters above 1000. MHz. I might note that the data to be displayed here provides engineers for the first time with significant component information in graphic form on capacitors at elevated frequencies.

Viewing pulse circuit and component problems from the frequency domain as well as from the time domain will often provide a key to solution of otherwise apparently difficult design problems.

Various network analyzers are available to determine the characteristics of any given two-port "black box". In our case, the "black box" consisted only of a capacitor across a gap on a 50 Ω microstrip line.

The measuring system used was an HP8542A, a computer controlled network analyzer. The parameters selected were the so-called "scattering-parameters" described in detail in the Hewlett-Packard application note #95.

S-Parameters are a measure of the power transmitted and reflected from some circuit component on a 50 Ω line.

They are more convenient to use than most other parameters because they do not require that voltages and currents be measured- a task which demands open and short circuits, difficult to achieve at 1000. MHz and up.

For a bilateral, passive device, such as a capacitor with a short path-length (for instance a cubic shape), only two of the four normally necessary parameters are required: the transmission coefficient S_{21}, and the reflection coefficient, S_{11}, (other capacitor physical configurations, such as oblong shapes, might require that all four be measured and provided.)

Some representative parameters are displayed in the following set of graphs, which are vector quantities displayed here as a magnitude and an angle. They could just as easily be displayed as two vector components and their signs (to show quandrant). They are shown up to 1 GHz, but we now have data to 3 GHz, and are working on testing for data up to frequencies equivalent to rise times of 50 ps.

When used as a DC-block on a transmission line, coefficients S_{21} and S_{11} may be used to determine signal attenuation in dB from: Insertion Loss = $10 \log_{10} (1/|S_{21}|^2)$, which for reciprocal devices may be separated into Dissipative Loss = $10 \log_{10}\left(\frac{1-|S_{11}|^2}{S_{21}{}^2}\right)$

and Reflection Loss = 10 $\log_{10}$ $(1 - |S_{11}|^2)$.

Note that the Transmission Loss versus Frequency graph (Figure 1), begins at 300. MHz. Below this, the reflected power due to the high capacitive reactance (of the values here initially chosen for measurement) dominates. The graph of S_{21} (magnitude) versus frequency is shown in Figure 2. An S_{21} magnitude of 1.00 signifies 100.% signal transmission, that is, no transmission loss, and therefore no pulse attenuation.

An S_{21} angle (Figure 3) of only a few degrees over a wide frequency range signifies a capacitor suitable for pulse transmission due to the low pulse shape distortion that would occur (thru differential phase shift of its various Fourier components.)

The reflection coefficient, S_{11} (magnitude), is equivalent to the microwave engineer's gamma (Γ), and from its magnitude may be determined (thru charts or calculations) the equivalent amount of power reflected before it ever could enter the capacitor to be transmitted.

Gamma is related to another quantity, commonly used when discussing transmission lines, called VSWR (Voltage Standing Wave Ratio). The three are all related as follows:

$$\text{VSWR} = \frac{1 + \Gamma}{1 - \Gamma} = \frac{1 + S_{11}}{1 - S_{11}}$$

The best possible VSWR is none at all, but this would require an S_{11} of zero, and that is nearly impossible to attain. The equivalent VSWR is displayed in Figure 4.

One practical significance is that it sets forth the ratio of load impedance to transmission line impedance, that is, for a 50 Ω line, a 25 Ω load, or a 100 Ω load would each have a VSWR of 2-to-1.

As many of you are familiar from time domain reflectometry in pulse testing of transmission lines, when a pulse signal is sent down a line and reaches an impedance different from that of the line, some of it is reflected and never gets a chance to be transmitted further.

This reflected power effectively subtracts from the original pulse amplitude, and is shown in Figure 5.

It may be readily solved for as shown in the front pages of the older editions of the "Microwave Engineer's Handbook and Buyers Guide" (Horizon House).

These techniques can be applied to any circuit interconnection or component, and where a very wideband DC-block is required on a transmission line, will tell in advance a great deal about the suitability of a particular choice and its exact characteristics over any given frequency range. Where load resistance levels are low, a capacitor used as a DC-block must have a very low equivalent-series-resistance, otherwise the transmission loss will be high. Conversely, a low S_{21} value often *implies* a high equivalent-series-resistance.

A capacitance value that is too small will produce a high VSWR and a high S_{11} magnitude at the lowest frequency equivalent of the pulse train.

A capacitor that is too inductive will have a large variation of the angle of the transmission coefficient versus frequency above the range where the VSWR had finished its normal original steep drop off to a low value.

Line-to-line coupling in a proposed pulse circuit layout will reveal itself beforehand by an increase in line Insertion Loss, or a decrease in the transmission coefficient of the line at the frequencies at which coupling occurs.

Thus, the techniques of the microwave engineer, if employed in pulse circuit design, can have a very great bearing upon the improvement possible in transmission of information at high data-rates. They provide him with a whole new set of eyes with which to see what is going on in his components and their interconnections.

BIBLIOGRAPHY

"Reflection and crosstalk in logic circuit interconnections" by J.A. DeFalco, IEEE Spectrum, July, 1970, pages 44-50.

"Transistor Pulse Amplifiers" by V.G.K. Reddi, Fairchild Semiconductor Corp., Application Note #APP-32.

"A 320. Mc Broadband Amplifier" by L.C. Drew, H.F. King, and A.G. Atwood, Solid State Design, Oct., 1964, pages 19-22.

"Digital Communications at Gigahertz Data Rates" by C.L. Cuccia, J.L. Spilker, and D.T. Magill, The Microwave Journal, Part I: Jan, 1970, pages 89-93; Part II: Feb., 1970, pages 87-92; Part III: April, 1970, pages 75-80.

"Germanium IC's point the way towards picosecond computers" by A. Reisman, Electronics, March 3, 1969, pages 88-93.

"Working with subnanosecond pulses" by E. Dilatush, EDN, March 1, 1970, pages 33-46.

"The use of digital circuits in data transmission" by P.J. Van Gerwen, Phillips Technical Review, Vol. 30, No. 3, 1969, pp.71-81.

"Coupled Transmission-Line Directional Couplers", by J.K. Shimizu, and E.M.T. Jones, IRE Trans. on MTT, Oct., 1958.

"Bandwidths Nomograph for Pulse Filter Network", by B.L. Adams, and T.M. Stout, Electronics Buyer's Guide, June, 1954, mid-month.

ADDITIONAL RECENT REFERENCES OF INTEREST

"Subnanosecond Switching and Ultra-Speed Data Communications by C.L. Cuccia, Data & Communications, November, 1971, pages. 19-25.

"Vector Voltmeter Measurement Techniques", PRD Electronics, Inc. Application Note. #22.

"Computer Researchers Achieve Their Fastest Switching So Far", NEWS section, Electronic Design #26, Dec. 23, 1971.

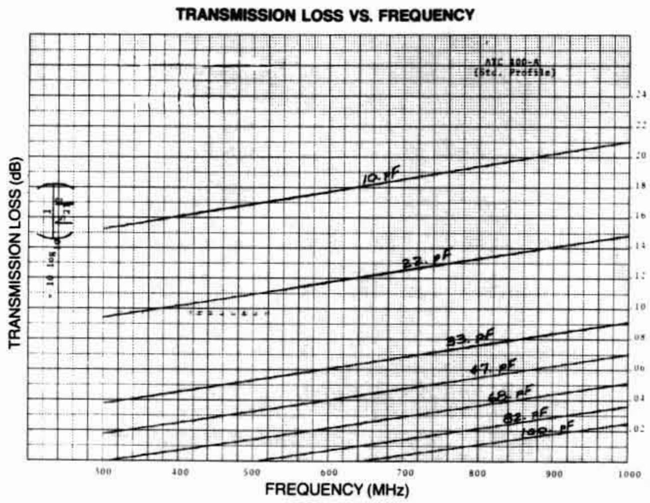


Fig. 1

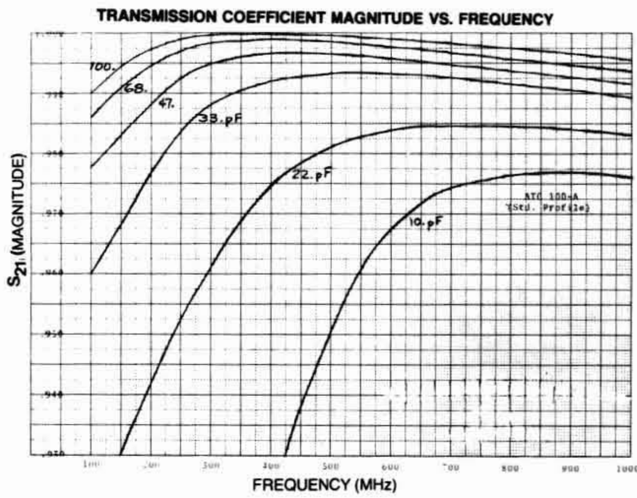


Fig. 2

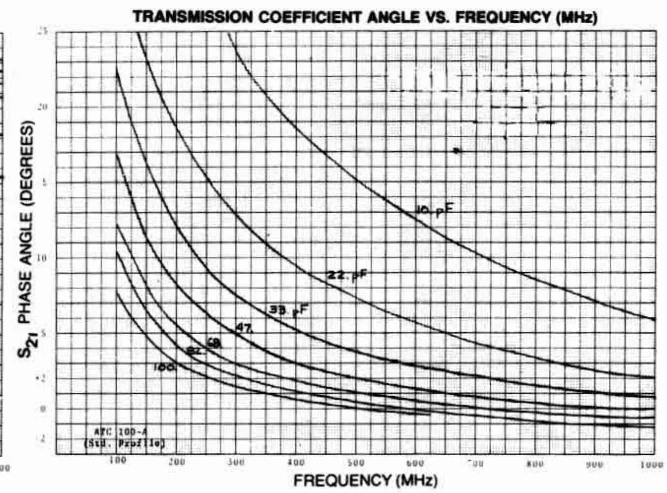


Fig. 3

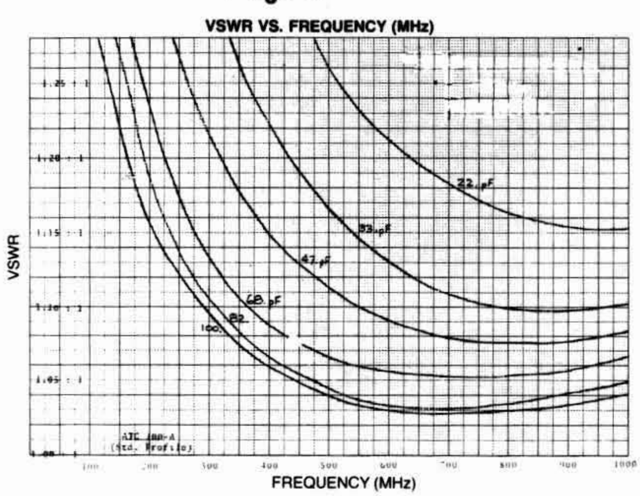


Fig. 4

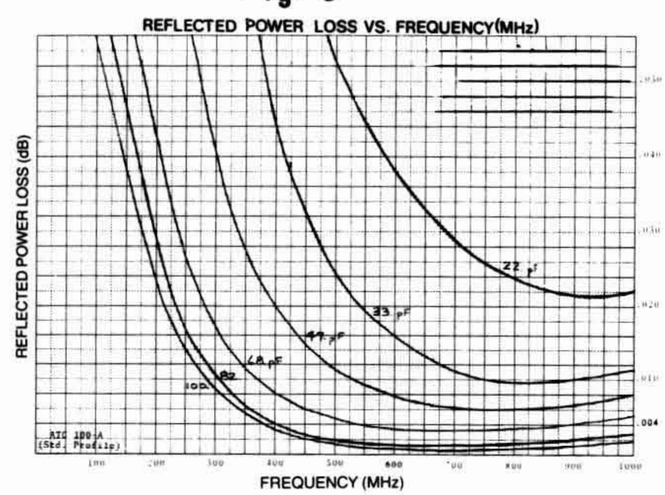


Fig. 5

CHIP CAPACITOR DIELECTRIC EFFECTS ON HYBRID MICROWAVE AMPLIFIER

Vincent F. Perna, Jr.
Vice President
Microwave Engineering

American Technical Ceramics
1 Norden Lane
Huntington Station, N.Y. 11746

ABSTRACT

To help the design engineer attain the full capability of his semiconductors, ATC provides: Test results of ATC's program of measurement of capacitor equivalent RF series resistance and resonance characteristics for various capacitance values from 100 MHz to 3 GHz.

RF transistor input current versus input power; current magnitude as a function of circuit bandwidth; power dissipated by the resulting circulating current.

Inter-relationship of Insertion Loss, unloaded Q and the S-Parameters.......with special emphasis on Q versus VSWR.

Delineation of transistor average minimum collector efficiency versus average power gain as functions of frequency, RF output power, and DC working voltage.

A study of the effects of interstage losses on driver junction temperature, and the resulting failure mechanisms of forward-bias, second-breakdown, aluminum/silicon mass transport, and the consequent expected transistor lifetime.

INTRODUCTION

The electrical characteristics of ceramic materials are strongly influenced by temperature, applied radio-frequency, and voltage. Their physical characteristics, as well, are affected by these influences.

These changes have a very direct bearing upon not only the installation method which should be employed when mounting chip capacitors in hybrid circuits, but also the gain, power output, bandwidth, and lifetime which may be expected from the completed circuit.

Due to the importance of this information to the design engineer, ATC offers in this paper: physical-effect data (particularly those of temperature on dimensional stability of ceramic-dielectric capacitors), resonance effects (both series and parallel), apparent-Q versus VSWR, more extensive unloaded-Q data, the related equivalent series resistance---which is really what the engineer working with power hybrids should know concerning capacitors---and, test data to 3 GHz on our ATC 100 microwave porcelain capacitors.

Using this data today, we will examine a typical power amplifier design in the several hundred MHz region, and discuss both the instantaneous failure and the slower "wear-out" mechanisms of transistor chips, plus, we will also examine circuit performance improvements that have been reported in hybrid amplifiers when using very low-loss, chip capacitors.

DISCUSSION

As many of you who have done wideband, matching-network designs for hybrid amplifiers know, we have been providing the S-Parameters, Insertion Loss, and VSWR of chip capacitors when used as a DC block on a 25 mil line on an alumina substrate. In addition to the series parameters, we will also be shortly providing the shunt parameters both in "Y" and S-Parameters at the end of a transmission line to ground. In addition, the unloaded capacitor Q, which will be provided, will be of a detail, as to cover the capacitance values typically used in such amplifiers for tuning purposes up to L-Band. For each of the capacitance values displayed from 10 pF to 50 pF, we will also be providing the equivalent series resistance of those capacitance values, both as functions of frequency and capacitance value. With this information, the hybrid circuit designer will be able to determine what his actual I^2R losses are, and thus determine, in advance, their influence on the gain and power output of his circuits. Other high-frequency data will shortly be provided up to X-Band on

As presented at the ISHM Conference, October 13, 1971

a new capacitor whose loss characteristics when used as a DC block have been found to be nearly indistinguishable from that of the line itself up to 11 GHz.

As you can see from Figure 1, which is our test data to 3 GHz on an ATC 100, Case A, 47 pF unit, even with a capacitor as large as a 50 mil cube across a gap of 20 mils in a 25 mil line, the VSWR is extremely low, even at S-Band. In the regions above L-Band one might begin to expect some transmission loss due to fringing in air and its interaction with the capacitor. However, as we can see from Figure 2, with higher dielectric constant substrates, fortunately most of it is still basically underneath the line. Consequently, for a hybrid amplifier, a capacitor chip of this particular size with internal metallization not greatly different from that of the line on the alumina substrate, will give very satisfactory performance. This may be seen from the Insertion Loss column, where typical loss is of the order of .05 dB to .03 dB throughout this region. Consequently, we have an almost negligible transmission reduction with microwave porcelain capacitors, even when including the effects of VSWR.

When examining the subject of capacitors for hybrid microelectronics, it is useful to gain some understanding of the effect of the various dielectric materials typically used in such capacitors. Here, for example, in Figure 3 is indicated a listing of the various types of ceramic materials commonly used and their equivalent dielectric constants and dissipation factors at a fixed frequency of 100 MHz. As it may be seen here, Barium Titanate-type materials (for example, those used in the so called high "K" dielectric capacitors), have a dielectric constant of the order of 8000. Unfortunately, it also has the disadvantage that it has a relatively high dissipation factor of 0.1 = DF. As we can see from progressing down the chart in the direction of decreasing dielectric constant, we have an improving dissipation factor. In fact, if we examine an equation in an article of August, 1970, in MicroWaves Magazine entitled, "A Guide to Judging Microwave Capacitors", the equation for the dissipation in capacitors in watts per cubic centimeter, that is, per unit volume, is given by: $W = E^2 \varepsilon_r \tan\delta\, f\, (55.5 \times 10^{-6})$, which tends to imply that the dielectric constant, loss tangent (or dissipation factor) and frequency are theoretically independent of one another. However, in real life, as we can see from this present tabulation, there is an approximate dependence of dissipation factor upon the magnitude of the dielectric constant. We see, by contrast, at the bottom of the chart, that this is not a hard and fast rule, since the ATC 100 Microwave porcelain has a somewhat higher dielectric constant than alumina, but it has a radically lower dissipation factor (by nearly an order of magnitude). What is important here is that the Watts-per-unit-volume of RF power dissipated in a hybrid amplifier will be dramatically lower when using microwave porcelain than with alumina, as a capacitor dielectric, and much more so than the commonly used NPO dielectric material.

It is this Watts-per-unit-volume of dissipation in a capacitor which is important to the hybrid circuit designer. It necessitates his examining carefully manufacturers' published data on both the high-power and low-power, unloaded capacitor-Q. (Q varies with temperature.) These data, if not available, may be tested for (see coming ATC Application Notes.)

Refering to Figure 4 for a moment, we can see that the material Barium Titanate, is commonly used in many forms of capacitor dielectric, all the way from the so-called high-"K" to the NPO, and that it has one very useful characteristic: you can get tremendous amounts of capacitance per unit volume with it. Unfortunately, it has some disadvantages that one should be aware of: it has detrimental effects on the temperature coefficient of capacitance, dissipation factor versus voltage, and physical modulus of extension versus temperature; it also exhibits piezoelectric effects and aging. In addition, it has another disadvantage of being very porous (internally and externally) plus having instability with temperature, frequency, time, and voltage, and, of being "lossy" when passing RF energy.

Let's examine some of these specific effects in order that we may better apply capacitors in hybrid amplifiers. In some cases this amplifier may wish to be unsealed or open to the atmosphere for reasons of cost. Dielectric porosity has the disadvantage that it allows atmospheric moisture and solvent-borne chemicals to penetrate the body of the capacitor. If you have a wet day you are going to experience deleterious effects. These penetrants can enhance internal migration of the metal electrodes under the influence of the applied electric field, and cause the capacitor to eventually short out, possibly causing destruction of active components on that substrate, as well.

All dielectrics vary in dielectric constant with temperature. If we take for example, the so called K1200 material, commonly noted as having a "BX" characteristic, we see from Figure 5 that it has a variation of dielectric constant, and therefore, capacitance, of around ±15% over the temperature range of -55°C to +125°C. In particular, we should note the peak on the right hand side, which for various manufacturers lies anywhere between 100 to 150°C. This is the so called Curie Peak (which, as we will see later, also has physical effects associated with it).

Progressing to the NPO material, we see that it is much more stable than the K1200 material, since it varies in temperature coefficient only ±30 PPM/°C. However, it has a disadvantage that, although it only goes negative and positive about zero with temperature, it does not follow the same path on return that it did under increasing temperature (due to its content of Barium Titanate). This instability, or hysterisis, otherwise known as imperfect retrace, may also occur with other parameters such as frequency and voltage.

NPO materials got their name from the fact that their dielectric constant varies alternately negatively and then positively with temperature. That is to say, if you want to start with any particular capacitance value (as "zero" reference) and if you heat the unit up, it will drop down a little bit in capacitance, then exceed the initial value some, then drop down again, typically lying within a slightly negative-going "cone", but basically always staying within ±30 PPM/°C. Although its temperature stability looks good, you should be aware of its hysterisis and imperfect retrace.

Aging is a characteristic of very high dielectric constant materials wherein, once the temperature associated with the Curie Peak is exceeded, the dielectric constant exhibits an immediate jump upwards of 10 to 15% in magnitude, and does not come back down immediately. In fact, it drifts down over a period of time described as "decade hours", such that, after the first 10 hours it may be only down 3% from the peak that it jumped up to, and after 10,000 hours it may only be down around 7%. In many applications this makes absolutely no difference; in coupling or by-passing it may actually be an advantage. However, you should not use such a material for a tuning application, or your circuit tuning would be varying all over the Smith Chart.

Comparing now the K1200 and NPO for their Dissipation Factor effects as functions of frequency, we see from Figure 6 that the K1200 material starts off at a moderate value of DF down in the audio frequency range of .01, but by the time it has gotten to 10 MHz it is starting to climb significantly, and by the time it is up to 100 MHz, the rate of increase of DF with frequency is so steep that when you get to L-Band it is of the order of 0.1 = DF. This means it has increased by a factor of 10 from its low frequency point, and the impact of this on an RF amplifier is seen in that, the unloaded capacitor Q is the ratio of the capacitive reactance of the equivalent series resistance (X_C / R_S = Q = 1/DF), and by having a high dissipation factor, you end up by having a low Q, which means a high equivalent series resistance. If we try to use this material in a tuning capacitor that is to have, say, 30 Ohms of reactance at L-Band, and it has a dissipation factor of 0.1, we end up with an equivalent series resistance in that capacitor of 3 Ohms. If we try to put any significant RF power thru such a capacitor, it would burn up because of the high temperature resulting from the high dissipated-Watts-per-unit-volume. It would have not only a poor low-power Q, but an even worse high-power Q, since capacitors when they dissipate RF power also degrade in Dissipation Factor and therefore in Q as well. The result is thermal run-away and self-destruction.

Examining now the effect on dissipation factor of increasing frequency in the NPO material, we note that it starts off at a moderate value of the order of .0001 = DF in the audio range, but once it gets up to 100 MHz it has risen rather significantly, and from there on up to L-Band the climb is very dramatic, also rising to a dissipation factor of about 0.1 at 1 GHz. This is roughly an increase by a factor of 1000. The disadvantage here again is that the equivalent series resistance of a capacitor is given by: R_S = (X_S DF) and if you tried to put any significant RF current thru it, it would get quite hot. To give you a feeling for how hot in practical terms, if we were dissipating only 40 milliwatts of power in a capacitor 0.1 inch on a side, we would have the same energy-density in Watts - per - unit - volume as the metallic portion of a 40 watt Ungar soldering iron. The reason it doesn't melt its terminals and fall off the board is that the heat removal capability of a microelectronic circuit on an alumina substrate is still sufficient to carry the heat off at a rate which will keep the temperature of the solder connection below its melting point. If you try, however, to put a large amount of RF current through this capacitor, its rate of heat generation will at some point exceed the rate of cooling available to the capacitor from the hybrid circuit, and it then *will* fall off the board. So, *both* NPO and K1200, while satisfactory for some applications at lower frequencies, degrade circuit performance from about 50. MHz on up.

Comparing the dissipation factor effects as a function of RMS voltage, we note that the K1200 material starts off at its common value of .01, but near the peak of every half cycle of RMS voltage (let's say we are talking about a Class-C RF amplifier now) we will find that the D.F. has jumped up a full order of magnitude or so. As we see from Figure 6, if we were to have RMS voltage swings of the order of 30 volts or so, we would easily have a times 10 increase over a portion of each half cycle. This results in heat generation pulses which contribute to the overall degradation of the amplifier because of the residual heat build-up inside the capacitor and its negative effect on Q.

NPO materials on the other hand are far better, having only a very slight effect, even for RMS voltages of the order of 35 or so.

If we now go to Figure 7 and examine the modulus of extension of Barium Titanate as a function of temperature, we will note that it has 4 crystalline phase states and 3 points of crystalline reorganization. When changing from one of these phase states to another with increasing or decreasing temperature, a physical reorganization in the crystalline structure of the dielectric material takes place and the physical extension that results externally is not smooth. If you were to take a metallic aluminum rod and heat it over a flame, you would be able to measure a quite smooth increase in its length. Conversely, if you were to put it in a refrigerator, you would note a rather dramatic, but smooth decrease in length. Capacitors containing Barium Titanate, however, exhibit abrupt changes in dimension when going from orthorombic to to the tetragonal, or from the tetragonal to the cubic, and it is up in this latter phase-change region in the area of the Curie Peak that you run into difficulty. Consequently, when the capacitor tries to expand in length as a function of temperature beyond the limits imposed by the solder preforms, it tends to buckle, and if you have ever tried bending a piece of chalk, you know it doesn't work very well. The result is that the capacitor tends to delaminate or crack internally, and you also find fractures at the end of the capacitor between the internal plates and the end metallization. These don't show up until you try to cool the capacitor down again, as you can see from the bottom of Figure 7. The ceramic tries to shrink, but its end metallization is contrained at the bond to the substrate, so the ceramic pulls away from the metallization. Since this bond was already fractured, the end result will be an open circuit, which can wipe out your active devices.

We feel that a cubic shape is more rugged than a long and slender format, since the longer the body, the greater the total dimensional change.

The industry has a lead-style designation for leadless units consisting of the word "chip". This means simply that the ceramic body has been silvered on its two end-faces. Sometimes they also use the term "solder-dipped chip" or "pre-tinned chip". We designate such a unit as a "pellet", which merely means it is a silvered chip whose ends have been dipped in molten solder. (Figure 8)

Fortunately for the ATC 100 microwave porcelain, we do not have problems with the thermal coefficient of expansion of our materials the way ceramic types do. The ATC 100 material has roughly the expansion coefficient of alumina, therefore no bond disturbance will occur, either when cooling down after running through a furnace, or mounting in a furnace on a hybrid microelectronic amplifier while the bond is still not solid, or, when being temperature cycled later.

There are a number of mounting methods. One commonly employed with chips is: the capacitor is mounted on solder preforms and then run through a furnace at such a temperature that the solder melts and adheres to both the capacitor and the substrate to cause a good bond. One common cause of failure here is poor silvering. The bond also has to rely, in some measure, upon the capillary action of the solder between the capacitor and the substrate. It also must rely upon surface tension to fight gravity on its way up the face of a capacitor to make a bond. As a result, it is common practice to check the bonds, after the chips have been furnace-mounted, by what some people call "tap-tap" testing, in which they try to dislodge the capacitor from the board to see whether or not there was a good bond formed. This can be an expensive labor-proposition. What we call pellets, by contrast, have the advantage that the solder is already all over the face of the capacitor, and when heated, just runs down to the substrate, causing a very obvious, large, fillet bond that does not require any manual "tap-tap" testing to see whether or not a good bond was formed. It does, however, have the disadvantage that pellets are more expensive than chips, so you have to trade off the cost of one versus the other for the most economical production situation.

Now, taking these various capacitors and mounting them on a transmission line, we can then measure them on a network analyzer for their relative performance versus frequency and RF power level. Refering to Figure 9, we see that,with a capacitor mounted as a DC block on a transmission line and some RF energy propagating down the line, when the RF comes to the gap in the line over which the capacitor is sitting, it has to decide what its going to do. Some of it takes the short, easy path through the first few layers of the capacitor, but much of the remainder has to climb up the face of the capacitor, go across near the top, then down the other side. This results in differing path lengths, and a phase shift among currents, which, at harmonically related frequencies, can result in Insertion Loss peaks. If mounted in shunt to ground, as at the bottom of Figure 9, a very narrow- band (high Q) increase in the resistive characteristics of the capacitor results. It does not stay out on the negligible-Ohms portion of the chart, but starts coming in towards the center-right, which is equivalent to an increase in the resistive characteristic. If this were the result of a high circulating current inside the physical body of the capacitor, this would imply there would be multiplied losses, because of the increased current traversal of the same path. This shows up as an increase in resistance. There is also an increase in the VSWR of the capacitor when mounted as a DC-block, so it tends to indicate that there is a parallel- tuned network which is reflecting power.

It struck me that this might be reduced, if, instead of having all the plates parallel to the surface of the substrate, they could be made vertical, and thus all of the current would reach all of the plates simultaneously. In doing this we found that this eliminated the first, third, fifth, and all odd-order resonances (note: all are approximately harmonically related), thus doubling the usable frequency range between these resonant points. These, incidently, occur generally above the frequency of series resonance, that is, after crossing from the lower (capacitive) to the upper (inductive)portion of the Smith Chart. This may come as a surprise to some people who are used to thinking only of <u>series</u> resonance in capacitors (probably a hang-over from the days of tubular capacitors, which were long and narrow in relation to their diameter).

All capacitors exhibit this effect, but the lossier a capacitor is, the less sharply defined these effects are. To broaden the bandwidth and reduce the amplitude of the effect, the price you must pay is greater loss at all frequencies. After considerable research, we have developed a capacitor, which, using new materials and new format, essentially avoids this problem up to 11 GHz or more. Until it is formally announced, you can partially avoid the problem by asking for "green dot" marking, which alerts the user to the option of vertical orientation of the internal electrodes. The green dot should be viewed somewhat as the notation on a packing crate which says "this end up". (Figure 10)

Taking now some of these capacitors that we have been talking about and examining their test data (particularly VSWR and loss) as functions of frequency, you will note in Figure 11 that an ATC 100-A "50 mil cube" with a 10 pF value when used as a DC block on a transmission line, has a negative gain, or rather, an insertion loss, at the lower frequencies that is fairly high. In fact the VSWR is *quite* high, and taking a look in particular at the area of 150 MHz, we note that we have a VSWR of nearly 6:1, and an Insertion Loss of almost 3. dB. Some engineers have noted this, and that the Insertion Loss decreases with increasing frequency, and they have felt that the Q at the lower frequencies must be pretty poor, but gets increasingly better as one goes up in frequency. From the mathematical description of capacitor Q in Figure 12, we see that Q is given as the reactance over the equivalent series resistance (X_C/R_S). We note from the bottom formula, that with frequency in the denominator we would anticipate that the Q would decrease with increasing frequency, and as you know from your measurements in the laboratory, this is exactly what happens. Consequently, we must look elsewhere for an explanation of this problem.

Refering to Figure 13, we note that, for increasing VSWR, we find an increasing reflected power and an increasing transmission loss, so by the time we get to a VSWR of 6:1 we have half our power reflected----- which is the 3 dB that was lost at the 150 MHz point. Thus, the capacitors do indeed have a very high Q. What is happening is that the power never gets into the capacitor to be dissipated. It's all being reflected, as one might expect from such very small values of capacitance as 10 pF at a low frequency. You are simply choking off the current flow.

Refering to Figure 14, we see the unloaded -component Q as a function of frequency in the frequency range of 100 MHz through a Gigahertz, and capacitance values typically used for tuning purposes over this frequency range. Although this data is what has traditionally been hoped for, especially in this detail, what is even more useful (Figure 15) is the equivalent series resistance of these same capacitance values over that frequency range. From this we may determine, from the input and circulating currents, what our I^2R losses are, and their effect on gain and power level. As we may note here from this graph, we will be talking about equivalent series resistances of the order of 12 to 40 milliohms.

From Figure 16, we can see a specific application of the information that we have been developing here. Into our hybrid amplifier there is a certain amount of RF input power. The transistor, let's assume, is in the frequency region of around 200 MHz, and thus might be expected to have an input reactance that is inductive. In fact, the inductance and base-emitter diode-resistance of this device may commonly be obtained from the data sheets of the semiconductor manufacturer as a function of frequency and power.

From this information, we can see (in the first equation) the loaded Q of the network that is available when using this transistor with a shunt capacitor to provide a single tuned network at a given frequency. We can also see that there is a certain bandwidth loaded circuit Q relationship, described in terms of an input-reactance to input-resistance ratio. We note also that the effective circuit input resistance is greater than that of the base-emitter diode alone. It is given by the loaded Q of this network squared, plus 1, times the base emitter diode resistance. This is the effective resistance into which the input power goes, and determines the input current to this single-tuned network (as given by the bottom equation).

If we assume we have a transistor with a base-emitter diode resistance of one Ohm and a loaded Q of 2(=50.% bandwidth), we would then have an effective input resistance at band center of:

$$R_{eff} = (Q_L^2+1)R_{in} = (4+1)1 = 5.\text{Ohms}.$$

If we further assume that we must put 15 Watts into this device, Figure 17 shows that we are forced to accept an RF input current of nearly 2. Amperes.

Turning now to Figure 18, we may note that the circulating current is given by the loaded Q of this single tuned network times the input current. Thus, the power dissipated in this first shunt, tuning capacitor is given by the circulating current squared, times the equivalent series resistance in that capacitor. The problem which now arises is to how to reduce this in order to reduce the wasted RF power evidenced as high heat per unit volume, and thus, not only decrease the temperature in the capacitor, but also increase the gain and power output of the network.

As you can see from the bottom equation, we can't very well reduce the input current, because that is a function of RF power, which is fixed. If we try to reduce the loaded-circuit Q, we affect the bandwidth and increase the cost. The only thing we can do to reduce the power dissipated in tuning capacitors is to reduce their equivalent series resistance. This requires a low-loss, microwave-porcelain dielectric.

Refering now to Figure 19 for a practical application of all this, let us assume that we have been given the job of devising a block diagram for a 400 MHz, 10 watt, RF amplifier in a hurry. Preliminarily, the design engineer may be asked whether or not he feels that such power output over a certain bandwidth is feasible, and, taking a quick look at some of the manufacturer's data, one would be led to the conclusion that "Yes, this is quite feasible". Now, in the questioner's mind you have been committed to 10 Watts over that bandwidth. Later you may learn that the additional requirements are: all components to be as cheap and as few in number as possible, low current drain, minimal heat sinking area, small size, and light weight; in addition, since the engineering staff has been reduced but the work has not, the design time is short. As a result, if the engineer goes to the bench to make a quick stab at seeing whether or not his prediction (now) can come true, and tries to follow the design requirements and use cheap components (high loss), the result will probably be 9 Watts out instead of 10. Although this may be disturbing, it is afterall, only 0.5 dB difference, and the engineer may feel he can pick this up somewhere without too much trouble. For example, he may feel that the driver transistor has enough gain capability that by increasing the bias he should be able to get out of difficulty.

Refering now to Figure 20, you will note that by increasing the driver stage bias, he lets himself in for problems in the areas of the efficiency of that stage, particularly when related to its thermal resistance. Figure 21, shows a study made of over 100 RF power transistors presently offered for sale in the output power region of 1 Watt to a little over 40 Watts, examining the average typical power gain and average typical minimum collector of efficiency as functions of frequency and collector operating voltage. Refering again to our 400 MHz example, we note that the collector efficiency of the output stage is going to be at least 55 or more percent, whereas the collector efficiency of the driver stage will be down somewhere around 40% or more ......roughly a 25% difference.

In discussing this problem with the transistor manufacturers, they pointed out that to solve their I^2R problems in the high current stages associated with high power, they have to optimize efficiency and let gain suffer. In the low-power region, where they don't have quite so much RF current to contend with, they can sacrifice efficiency and optimize gain. These, plus the other trade-offs which they are forced to live with in the real world, result in the problem of how to get rid of heat generated by this poorer efficiency in the driver stage.

As we can see from Figure 22, if we have a quoted minimum efficiency of 40% or better, and wish to correct for the 0.5 dB loss in gain by increasing the current in the driver stage, the power that was lost in those inefficient capacitors has a larger effect than we expected because of the poor efficiency of the driver stage. That 0.5 dB requires about 250 milliwatts more of DC power thru the driver, and while it is true that we would get our 100 milliwatts of RF out, the problem is that we will leave behind 150 milliwatts dissipated as heat.

If we are to abide by the design considerations laid down and use as few and as cheap components as possible, we would be operating most of them up toward the maximum portion of their capability. If we try to do this with transistors, we had better keep the maximum collector junction temperature below 200°C. Some people recommend a 10% safety margin, and operate at 180°C or below.

The effect of this type of false economy shows up in that the average thermal resistance of low-power devices in the 1 to 5 Watt output region is more than 24°C-per-Watt, and the 150 milliwatts left as heat, times 24°C/W, gives an increase of 4°C over the 180°C that we desired. We have just lost 20% of our safety margin. Well, although that doesn't seem like much, you had better start considering potential failure mechanisms.

In Figure 23, we note that we will have to face not only the well known instantaneous failure mechanism of Second Breakdown, but also the slower wear-out mechanisms of: mass transport of aluminum and silicon, etch pits forming in the silicon, and, swelling aluminum at the confluences of the emitter-stripes.

From Figure 24, we note that, if we had an infra-red scanning microscope and could scan across the transistor chip mounted on the substrate when putting RF power through it, we would indeed have an average junction temperature of the order of 180°C. We would also note that there would be areas that are cooler than that, and some that are much hotter than that, for as good as the transistor manufacturer is, he still has the real world to contend with, and there are always slight imperfections in window etching, metal deposition, processing of silicon, emitter balancing resistors, and so forth, and he ends up with having a percentage of devices which don't perform optimally. In fact according to the manufacturers, there can be temperatures on that chip up to 200% of the average junction temperature. Now, if we do get temperatures that high, we will surely worry about the integrity of the transistor, since if we inadvertantly augment the temperature of a hot spot to the point where it melts, it will no longer act as a semiconductor, but rather as a metallic conductor, destroying the operation of that portion of the device.

Let's say however, that the transistor we are using is not really poor, but only "marginally defective". Individual sections of the transistor chip may then be operating at high, but not immediately damaging, temperatures.

Refering now to Figure 25, we note that if we were to put a current typical of RF power transistors thru a wire 1 mil in cross-section, and examine the current density, we might have a moderate value of the order of 10,000 Amperes per square centimeter. By contrast, if we were to examine the tiny metallized inter-connections on the surface of the silicon, we might note current densities of the order of a million Amperes per square centimeter, since we are now effectively dealing with Angstroms of conductor thickness.

The problem described in the literature is this: an "electron wind" results, which increases in force with the square of the current density. Thus, a small increase in current results in a large increase in electron wind force. This means that if you were to have a narrow section of metallization, the current flowing would tend to have a scouring effect on the path through which it is flowing. For those of you who have been out in a rain storm in a western desert, this may recall that when the water comes roaring through down those gullies, it literally rips up trees, bushes, and bolders, and carries them along until it finally dumps its burden when the path widens and the water slows its rate of flow. We have an analogous condition inside the transistor. Slight imperfections in the metallization through the silicon dioxide windows form cusps and necked-down portions, and the resulting high current density tends to strip away the aluminum as a function of time and temperature.

When operating a transistor, we are not always allowed the most favorable environment. Often we may face an external ambient temperature of 125°C, which contributes to the internal temperature of the transistor. This may raise it to a point where we are now imparting significant Kinetic energy in the form of heat to the lattice atoms, and some begin vibrating somewhat free of the lattice. The result is, as we can see from Figure 26, that the electron wind roaring through these conductors carries away atoms torn free by impact. In addition, we have the property of silicon that it tends to dissolve into the aluminum, and the silicon may now be scoured away by the electron wind. Furthermore, aluminum tends to fill any vacancy where silicon was, thus, aluminum-filled etch pits start developing in the silicon and work their way down, eventually shorting out that section of the device. This is the first of the failure mechanisms mentioned under "wear-out" of transistors.

This scouring mechanism can be more readily grasped by looking at the bottom of Figure 26. Let's say we have got 1 aluminum atom which has been agitated by this input energy in the form of heat and readily scoured away. This causes the current density in its former neighbors on either side of it to increase. Thus they are more readily scoured away than they were before, and this process increases rapidly to the point where a gap can develop in that conductor. We will thus have a void in the metal path which will cause the transistor to cease operation in that emitter stripe.

The third mechanism that we have to contend with is similar to that at the delta mouth of a river, where now, down stream, this electron wind is carrying its burden of silicon and aluminum along and it comes to the point where the current density drops because the conductor widens, and so it unloads this "sediment". This starts forming humps and hillocks on the surface, and works its way upwards in the form of lumps and crystals and other irregularities. Since RF current travels mostly on the surface, it begins finding an extremely rough, long path. In fact, the literature notes that in this area of swelling aluminum, we can experience a path resistance of up to 200 times that of the normal emitter stripe. If this occurs, the transistor will have so much resistance in this area that it cannot effectively perform its desired function, and performance seriously declines in that area.

So, there are several mechanisms all working competitively to destroy transistor action on a long term basis as functions of current and temperature. The mechanism that usually wins is swelling aluminum. Thus, as a function of time, your transistor just slowly dies, and you may not know why.

One of the ways to avoid this problem is to reduce the amount of current that you are pushing thru the transistor by using low-loss, interstage-network coupling-capacitors, which allow you to drive the transistor less hard because now you don't have losses that you have to overcome.

Let us now examine what effect that extra 4°C at the collector junction has in terms of transistor lifetime.

In Figure 27, we see the relative lifetime of a transistor as a function of collector junction temperature. Semiconductor applications engineers have a rule of thumb that, for every 10°C you can stay below 200°C you will have double the device lifetime you would have had at 200°C. Thus, by operating at 180°C, you will have 4 times the lifetime we would have had at 200°C. The effect, then, of our 4°C increase due to lossy capacitors, is to cut our relative transistor lifetime by 25%. We can see also from this chart why transistor manufacturers recommend that for a good, long-term reliability the collector junction should not operate higher than around 140°C. Unfortunately, in the real world, we may have conditions which require us to operate in an ambient temperature of 125°C and this means that we would have a restricted output power because of the basic efficiency of the transistor. An alternative, of course, is to buy a higher power device with lower thermal-resistance. However, this typically means higher cost, lower gain, and changed bandwidth capability. Consequently, we often have to operate higher than 140°C.

To keep the transistors at the lowest possible temperature, we have to have the lowest possible RF current flowing through them. To do this, we have to have the minimum amount of unnecessary circuit losses. So, to have a good transistor MTBF (and thereby system MTBF), we must not waste power in the interstage networks. Wise economy, therefore, dictates that one examine the relative cost of low-loss versus high-loss capacitors in comparison to the cost of the power transistors needed in those two cases.

In Figure 28, we can see some improvements reported by engineers who, having examined the loss characteristics of various capacitors, have employed low-loss microwave porcelain instead of using the commonly chosen NPO material. We find from these examples (all of which are from different manufacturers) that, in case number one, power output was doubled from 7 Watts to 14 Watts simply by replacing NPO capacitors with microwave porcelain. In the second case, there was a bandwidth increase of 25% due to the elimination of the passband droop due to unnecessary losses at the high end. (As we noted earlier, capacitors tend to increase in loss as frequency rises.)

In the next case, by replacing lossy NPO capacitors with microwave porcelain, the DC current drain of a hand-held radio was reduced to the extent that instead of requiring 20 Watts of DC to get 10 Watts of RF output, only 13 Watts of DC were required. This dramatic improvement in the battery requirement allowed the manufacturer to offer their customer several options: a longer lifetime under normal operating conditions, a greater percentage of the operating time in the high-current mode of transmitting, a lower-cost battery, a smaller battery, or a lighter weight battery.

In a fourth case, there was a requirement for reducing the physical size of the high-power phase-shifters in a phased array radar from about 10 feet long, 4 feet wide, and 1/2 foot thick (each) down to the more managable proportions of a pack of cigarettes. With a very large klystron generating multi-kilowatts of power to a corporate feed which then divided this down to the individual array ports, the physical volume of all the phase shifters necessary for the phased array filled a rather large room. This was not what you would call a portable system. This approach had been forced upon them, since, to handle of the order of 1000 volts of RF, and currents of about 10 Amperes, they had to make it large enough to be low loss and not suffer dielectric break-down.

The desire was to shrink this using a thick alumina substrate, and ATC was asked to develop a capacitor which would handle a several kilowatts of pulsed RF power at the 50 Ohm level (several hundred Watts of average power). We not only designed this capacitor, but improved it so it could handle 8 kilowatts and then 15 kilowatts. This is equivalent to over 900 watts of CW at many hundreds of Megahertz flowing thru a capacitor with a physical volume of .001 cubic inches. However, sensing that this was not the limit of the capability of these low-loss, microwave-porcelain capacitors, the customer then connected 4 of them in a series-parallel quad and proceded to push over 40 kilowatts of RF power through them.

CONCLUSION

Where impedance levels are low or RF current high, any equivalent series resistance in a tuning or matching capacitor acts as an unwanted power divider and degrades gain, power output, DC-to-RF conversion efficiency, and circuit MTBF.

Greatly improved performance and transistor lifetime may be readily attained by simply replacing lossy capacitors by ultra-low-loss microwave porcelain capacitors.

ABOUT THE AUTHOR

Prior to joining ATC in 1968, Mr. Perna had been a consultant on microwave amplifier and multiplier design, particularly with East Coast companies, among them RCA, Raytheon, and IBM.

His original interest in electronics developed in the early 1950's and he soon became a licensed Radio Amateur. This developed into working as an electronics technician and subsequently as an engineer before and during his study at UCLA.

After graduation in 1960 with a Bachelor of Science degree in Applied Physics (Electronics), he developed a preference for wideband video, and later, RF amplifier and multiplier design.

One employer, Canoga Electronics, for whom he worked, was an early advocate of computer-aided circuit-design, and offered the engineering staff eight weeks of afternoon/evening instruction at the IBM facility. Mr. Perna took this opportunity, and Canoga further financed an additional eight weeks of special investigation by him into the problems of computer-assisted, high-power, varactor-multiplier design.

Later, during other company-sponsored evening-education, he was invited to deliver a paper on frequency multipliers before the Los Angeles IEEE Chapter on Circuit Theory.

These events were instrumental in his being invited East by RCA to engage in design of power "amplifier/multipliers" and "oscillator/multipliers" using the pumped C_{ob} of overlay-transistors.

During his association with several eastern corporations to design RF power amplifiers (to 2. GHz) and step-recovery-diode multipliers (to X-Band), he was asked to deliver papers on RF circuit design before their engineering staffs, and then, to offer staff training in design techniques.

Later, while providing design services to another client on high-power, wideband, VHF-thru-L-Band noise jammers, he became acquainted with the unique characteristics of ATC capacitors.

As a result of outlining to ATC the need for, and advantages of, supplying the engineering community with specific design aids, he was invited to join ATC to manage their Sales and Applications Engineering Departments. Since then, he has lectured extensively and written numerous articles for ATC, explaining the impact of low-loss, microwave capacitors on amplifier and multiplier performance.

REFERENCES

1.) "Electromigration Failure Modes in Aluminum Metallization for Semiconductor Devices", by James R. Black, Proceedings of IEEE, Vol. 57, No. 9, September, 1969, Pages 1587-1594.

2.) "Some Observations on Electromigration in Aluminum Films", by P.B. Ghate, Semiconductor Research and Development Lab., Texas Instruments, Inc., Dallas, Texas, Applied Physics Letters, Volume 11, No.1, July 1, 1967, pages 14-16.

3.) "Aluminum Conductor Failure", by J.R. Black, Proc. 3rd Internatl. Congr. Microelectron (Munich, November, 1968), pages 141-162.

4.) "Electromigration---A Brief Survey and Some Recent Results", by James R. Black, IEEE Transactions on Electron Devices, Vol. ED-16, No. 4, April, 1969, pages 338-347.

5.) "Electromigration in Thin Gold Films", by Thomas E. Hartman and James C. Blair, IEEE Transactions on Electron Devices, Vol ED-16, No. 4, April 1969, pages 407-410.

6.) "Aluminum Electromigration in Long Stripes", by J.M. Attardo, A.H. Landzberg, W.E. Reese, and G.T. Wenning, presented at the 7th Annual Reliability Phys. Symp. Washington D.C., Dec. 2-4, 1968, (pages 7 and 8 of Proc.).

7.) "Degradation of Evaporated Aluminum Contacts on Silicon Planar Transistors", by C.B. Oliver, presented at the 6th Annual Symp. Reliability Phys., IEEE Cat. 7-15C58, pages 209-215, Nov. 1967.

8.) "The Failure of Thin Aluminum Current-Carrying Strips on Oxidized Silicon", Ilan A. Blech and Harry Sello, presented at the 5th Annual Symp. Phys. Failure Electron., Columbus, Ohio Nov., 1966, (Proc. pages 496-505).

9.) "Electromigration in Metal Film Conductors", by Walter E. Mutter, presented at the Electrochem. Soc. Spring Meeting, Dallas, Texas, Abstract 62; pages 96-98, May, 1967.

10.) "Mass Transport of Aluminum by Momentum Exchange With Conducting Electrons", by James R. Black, Proc. 1967 Annual Symp. Reliability Phys., IEEE Cat. 7-15C58, Nov. 1967, Pages 148-149.

11.) "RF Power Transistor Metallization Failure", by James R. Black IEEE Transaction on Electron Devices, Vol. ED-17, No. 9 September, 1970, pages 800-803.

12.) "Short Pulse Rating of Integrated Circuit Conductors" by L.J. Kroko, Microelectronics and Reliability, Pergamon Press, 1970, Vol. 9, pp. 503-510.

13.) "Deposited Oxide Contours in Multi-Layer Metal Circuitry" by D.L. Tolliver and C.J. Santoro, Solid State Technology, April, 1971, pages 32-36.

14.) IEEE Transactions on MTT, July, 1971, Vol. 19, No. 7, pages 609-616.

15.) "GHz Transistor Destroys Itself at UHF", by Robert H. Cushman, EDN, Sept.1, 1970, page 59.

16.) "Power for Portables" by Mr. Robert F. Chase, IEEE Transactions on Vehicular Technology, Vol. VT-19, No. 4, November, 1970, pages 248-251.

17.) "Design Interface", EDN/EEE, May 15, 1971, page 58.

18.) Sanders Associates, Inc. Microwave Division's insert in Microwave Systems News, May/June 1971, Vol. 1, No. 5, pp.24.

19.) Microwave Filters, Impedance-Matching Networks, and Coupling Structures, by G.L. Matthaei, L. Young, and E.M.T. Jones, McGraw-Hill, 1964, page 87.

20.) "A New Look at Microwave Silicon Technology" by E.J. Rice, R.D. Gromer, K.M. Finn, B.J. Tilley, and W.E. Schaub, The Microwave Journal, February, 1969, pages 80-86.

21.) "Microstrip Amplifiers Can Be Simple" by George D. O'Clock, Jr., Electronic Design 14, July 8, 1971, pages 66-68.

22.) The Radio Engineers Handbook, by Keith Henney, McGraw-Hill, 3rd Edition, 1941, page 140.

23.) Communication Engineering, by William L. Everitt, Ph.D, McGraw Hill, 2nd Edition, 1937, page 575.

24.) Electronic Engineers Handbook, by R.W. Landee, D.C. Davis, and A.P. Albrecht, McGraw Hill, 1957, pages 4-30.

25.) "Evaluate Transistor Bandwidths the Easy Way", by V.F. Perna, Jr., Electronic Design Magazine 26, December 20, 1970, pages 40-42.

26.) "A Guide to Judging Microwave Capacitors", by V.F. Perna, Jr., Micro-Waves Magazine, August, 1970, pages 40-42.

27.) Vitramon Inc., Data Sheet P10D

28.) Radio Materials Co., 1970 Catalog

29.) San Fernando Electric Mfg. Co. Bulletin MC 031269C.

30.) Republic Electronics, Catalog B-1, (3/66), and, 1966 Catalog E-1.

31.) Monolithic Dielectrics, Inc., Bulletin No. 123.

32.) U.S. Capacitor Corp., Bulletin (no reference number).

33.) San Fernando Electric Mfg. Corp. Bulletin 31771-CC

34.) USCC/Centralab, Catalog (no reference number).

35.) U.S. Capacitor Corp., "Application Notes" (no reference number).

36.) San Fernando Electric Mfg. Co., Handbook #H31570

37.) "Precision Solid State Film Transport" by Dr. Charles M. Davis, Teledyne-Ryan Report #AFAL-TR-71-91, April, 1971.

FIGS. 1 - 28

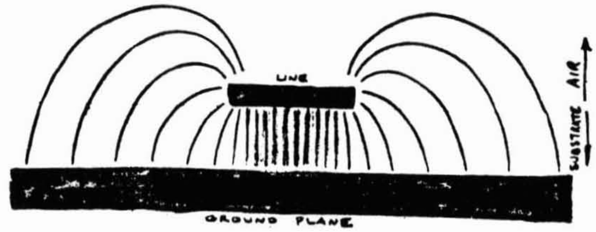
ATC 100-A-470-J-C-50-SP

FREQ.	VSWR	I.L.	Re S_{11}	Re S_{21}	Im S_{21}
500.	1.16	.03	-.07	1.00	.06
550.	1.15	.06	-.07	.99	.05
600.	1.14	.07	-.06	.99	.05
650	1.13	.06	-.06	.99	.04
700.	1.12	.07	-.05	.99	.04
750	1.12	.04	-.05	1.00	.03
800	1.11	.06	-.04	.99	.03
850.	1.11	.04	-.03	1.00	.03
900.	1.10	.01	-.02	1.00	.02
950.	1.10	.04	-.02	1.00	.02
1000.	1.10	.08	-.01	.99	.02
1100.	1.09	.06	.00	.99	.01
1200.	1.09	.07	.01	.99	.01
1300.	1.08	.03	.02	1.00	.00
1400.	1.08	.03	.03	1.00	.00
1500.	1.07	.07	.03	.99	.00
1600.	1.07	.08	.03	.99	.00
1700.	1.06	.07	.03	.99	-.01
1800.	1.06	.03	.02	1.00	-.01
1900.	1.05	.03	.02	1.00	-.01
2000.	1.05	.06	.01	.99	-.01
2100.	1.05	.06	.00	.99	-.02
2200.	1.04	.05	-.01	.99	-.02
2300.	1.04	.03	-.01	1.00	-.02
2400.	1.03	.07	-.01	.99	-.02
2500.	1.03	.08	-.01	.99	-.03
2600.	1.02	.08	-.01	.99	-.03
2700.	1.01	.05	-.01	.99	-.03
2800.	1.01	.05	.00	.99	-.03
2900.	1.01	.06	.00	.99	-.03
3000.	1.02	.08	.01	.99	-.04

FIG. 1

MICROSTRIP LINE
(CROSS-SECTION)

GENERALIZED FIELD DISTRIBUTION



REF: "USING STRIP TRANSMISSION LINE TO DESIGN MICROWAVE CIRCUITS" BY J.R. DANGL AND K.P. STEELE, ELECTRONICS, FEB. 7, 1966, PAGES 72-83

FIG. 2

DIELECTRIC MATERIALS COMPARED

TYPE	$\epsilon_r = K$	D.F. @ 100 MHz
$BaTiO_3$ ("HI-K")	8000	0.1
$BaTiO_3$	1200	0.03
NPO	30	0.002
ALUMINA	9	0.0005
ATC/100	15	0.00007

FIG. 3

DIELECTRICS

BARIUM TITANATE ($BaTiO_3$):

WHERE ENCOUNTERED:

- "K1200"
- "NPO"

EFFECTS:

- TC
- DF - VS - VOLTAGE
- MODULUS OF EXTENSION
- PIEZOELECTRIC
- AGING

ADVANTAGES:

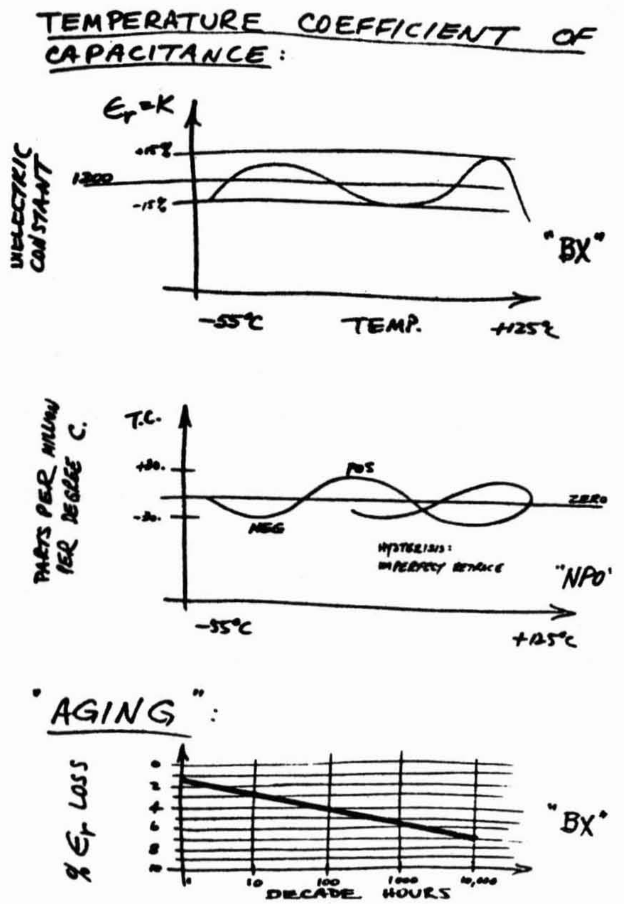
- PACKAGING DENSITY

DISADVANTAGES:

- INSTABILITY
- LOSSY
- POROUS

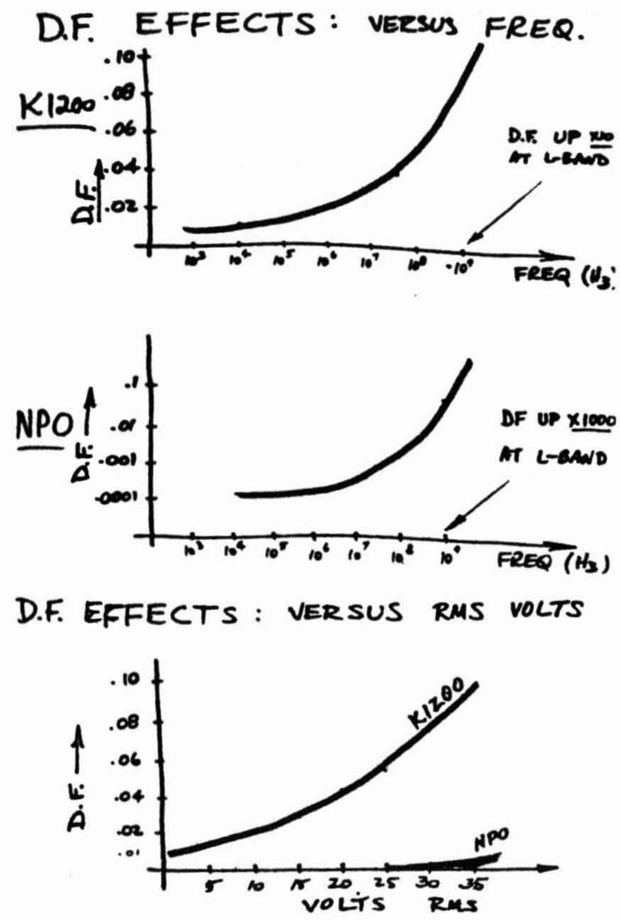
FIG. 4

CHIP CAPACITOR DIELECTRIC EFFECTS



TEMPERATURE COEFFICIENT OF CAPACITANCE:
DIELECTRIC CONSTANT
"BX"
TEMP.
-55°C
+125°C
PARTS PER MILLION PER DEGREE C.
T.C.
ZERO
POS
NEG
HYSTERESIS: IMPERFECT RETRACE
"NPO"
"AGING":
% LOSS
DECADE HOURS

FIG. 5



D.F. EFFECTS: VERSUS FREQ.
K1200
D.F.
D.F. UP X10 AT L-BAND
FREQ (Hz)
NPO
DF UP X1000 AT L-BAND
FREQ (Hz)
D.F. EFFECTS: VERSUS RMS VOLTS
K1200
NPO
D.F.
VOLTS RMS

FIG. 6

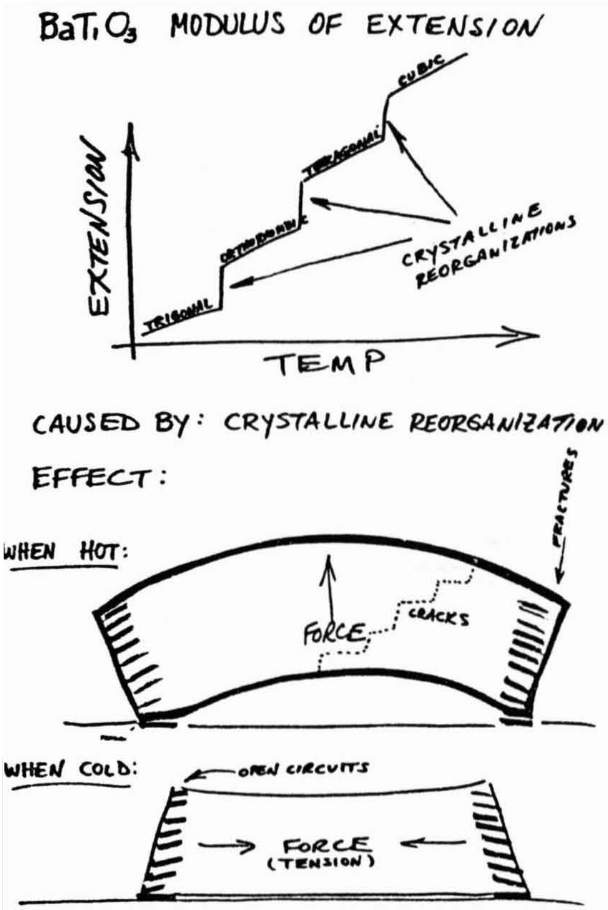


FIG. 7

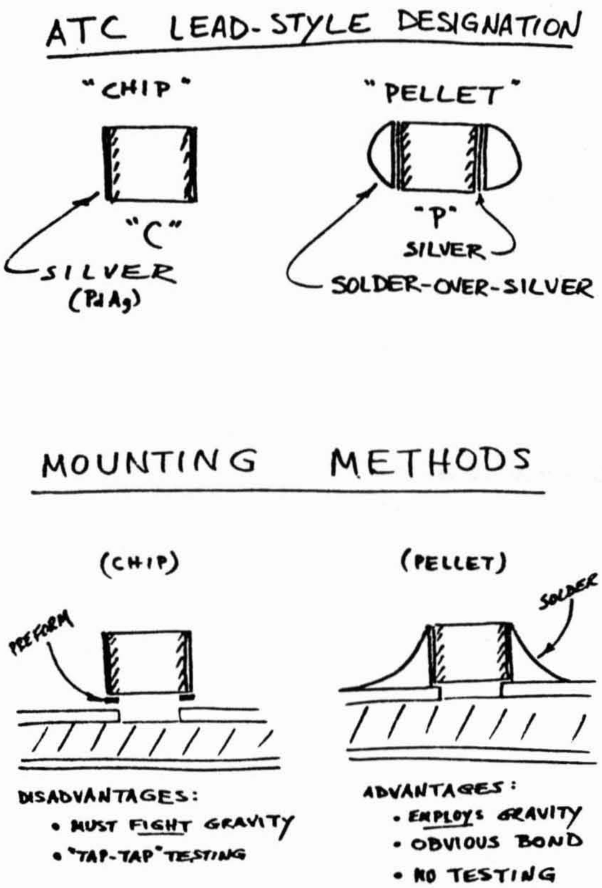


FIG. 8

CHIP CAPACITOR DIELECTRIC EFFECTS

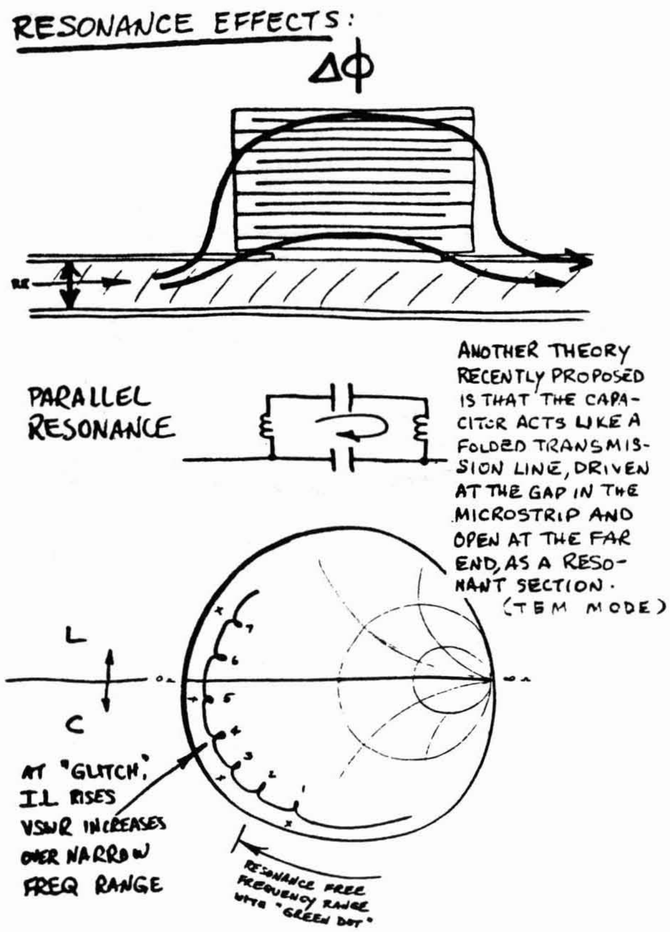

RESONANCE EFFECTS:
Δφ
PARALLEL RESONANCE
ANOTHER THEORY RECENTLY PROPOSED IS THAT THE CAPACITOR ACTS LIKE A FOLDED TRANSMISSION LINE, DRIVEN AT THE GAP IN THE MICROSTRIP AND OPEN AT THE FAR END, AS A RESONANT SECTION. (TEM MODE)
L
C
AT "GLITCH", I.L RISES VSWR INCREASES OVER NARROW FREQ RANGE
RESONANCE FREE FREQUENCY RANGE WITH "GREEN DOT"

FIG. 9

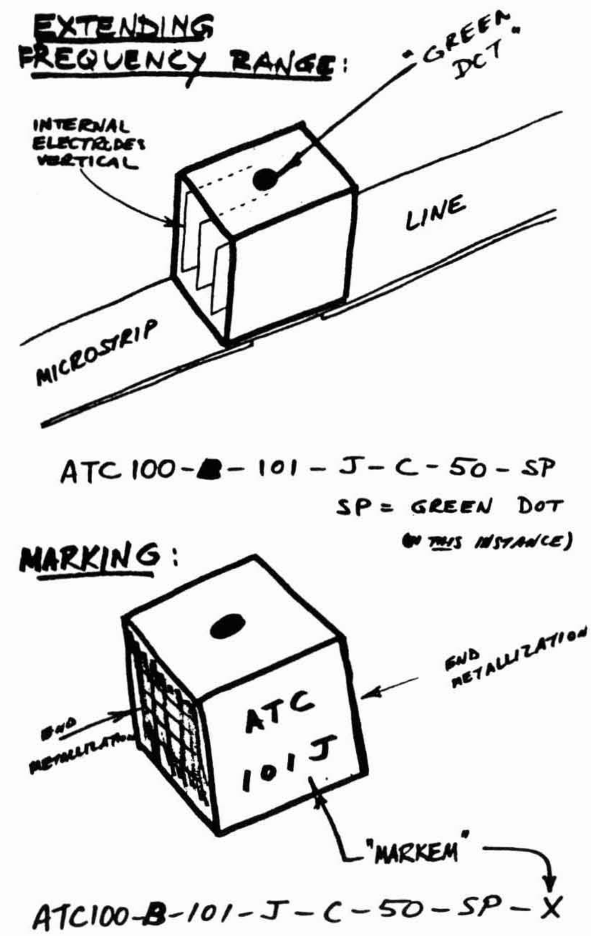

EXTENDING FREQUENCY RANGE:
"GREEN DOT"
INTERNAL ELECTRODES VERTICAL
LINE
MICROSTRIP
ATC 100-B-101-J-C-50-SP
SP = GREEN DOT
(IN THIS INSTANCE)
MARKING:
END METALLIZATION
END METALLIZATION
ATC
101J
"MARKEM"
ATC100-B-101-J-C-50-SP-X

FIG. 10

NETWORK ANALYZER
TEST DATA

ATC 100-A-100-J-C-50

	FORWARD	
FREQ	VSWR	GAIN
100.000	11.513	-5.24
150.000	5.960	-2.61
200.000	3.962	-2.16
250.000	3.039	-1.16
300.000	2.547	-1.07
350.000	2.244	-.89
400.000	2.037	-.66
450.000	1.883	-.72
500.000	1.772	-.51
550.000	1.685	-.45
600.000	1.617	-.28
650.000	1.554	-.24
700.000	1.501	-.23
750.000	1.458	-.22
800.000	1.424	-.27
850.000	1.397	-.21
900.000	1.372	-.23
950.000	1.355	-.18
1000.000	1.338	-.18

FIG. 11

$$Q_{unloaded} = \frac{X_c}{R}$$

$$Q_u = \frac{1}{2\pi f R C}$$

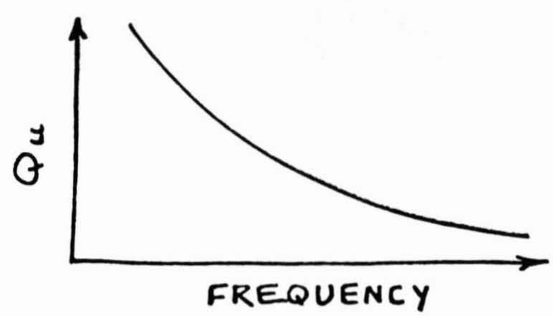


FIG. 12

CHIP CAPACITOR DIELECTRIC EFFECTS

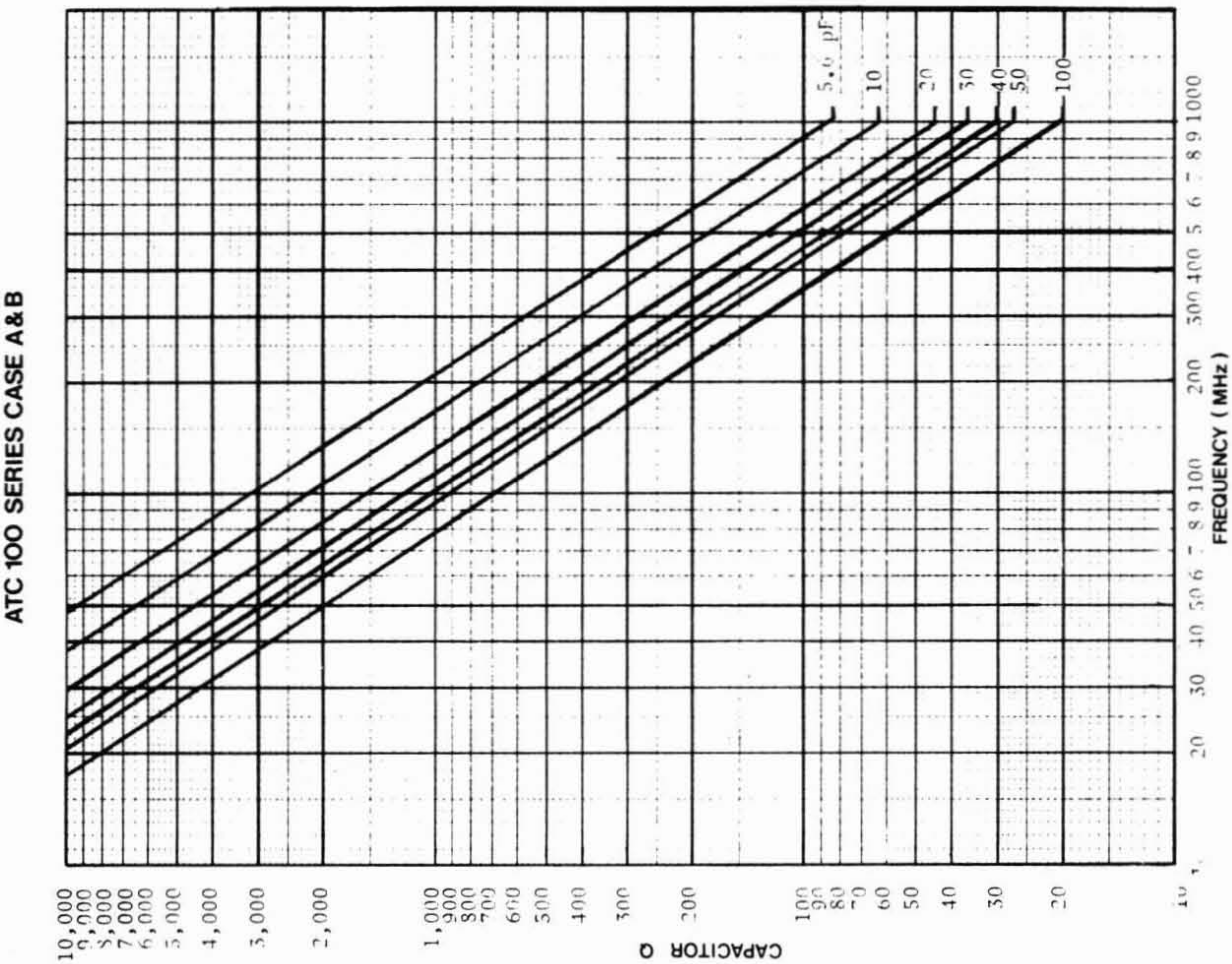

ATC 100 SERIES CASE A&B
CAPACITOR Q
FREQUENCY (MHz)
10,000
9,000
8,000
7,000
6,000
5,000
4,000
3,000
2,000
1,000
900
800
700
600
500
400
300
200
100
90
80
70
60
50
40
30
20
10
20 30 40 50 6 7 8 9 100 200 300 400 5 6 7 8 9 1000
5.6 pF
10
20
30
40
50
100

FIG. 14

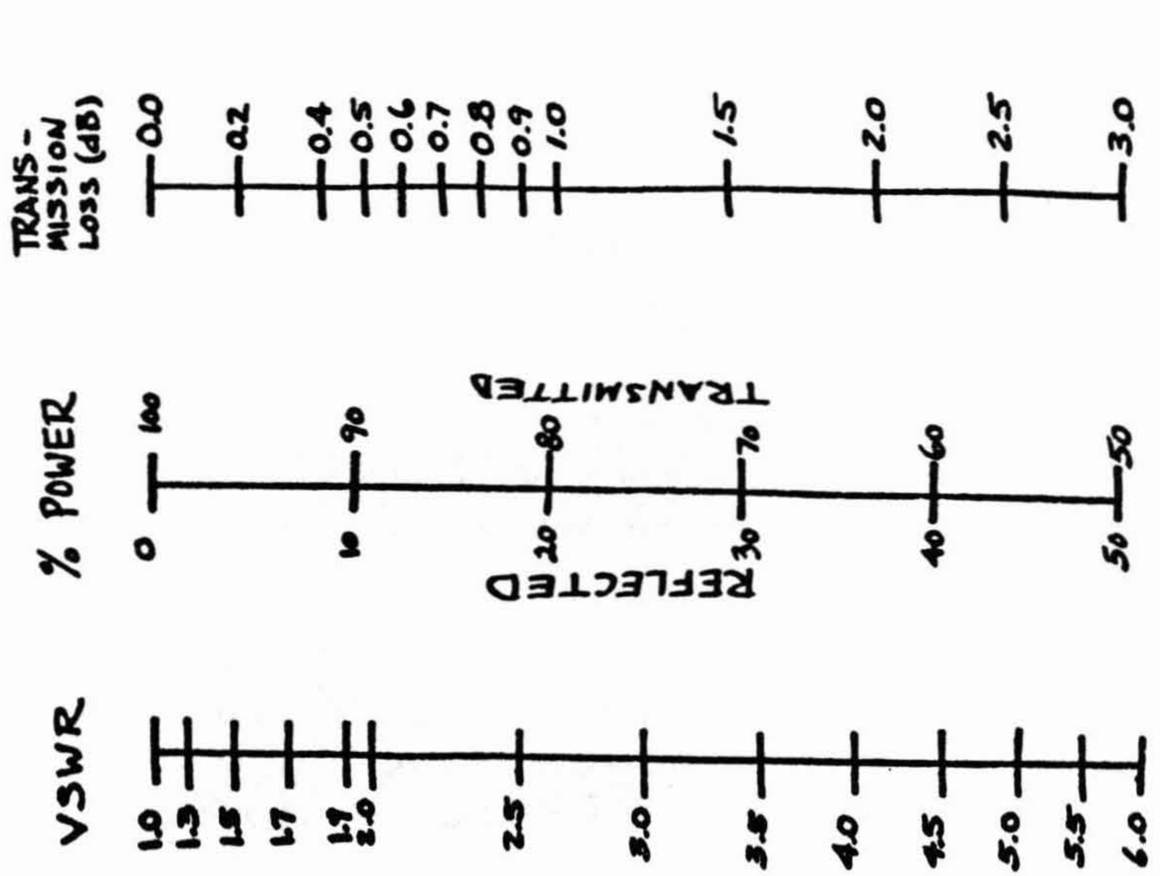

VSWR
1.0
1.3
1.5
1.7
1.9
2.0
2.5
3.0
3.5
4.0
4.5
5.0
5.5
6.0
% POWER
REFLECTED
0
10
20
30
40
50
TRANSMITTED
100
90
80
70
60
50
TRANS-MISSION LOSS (dB)
0.0
0.2
0.4
0.5
0.6
0.7
0.8
0.9
1.0
1.5
2.0
2.5
3.0

FIG. 13

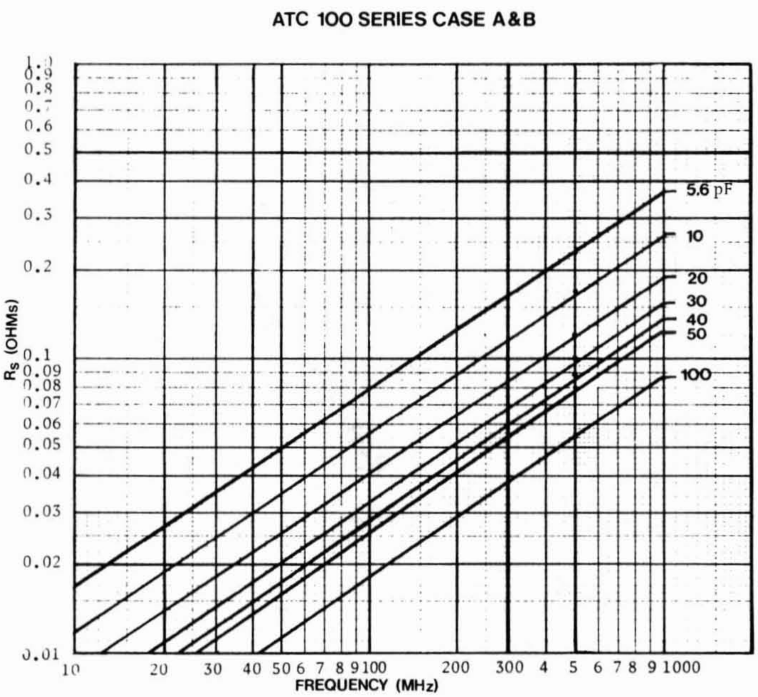
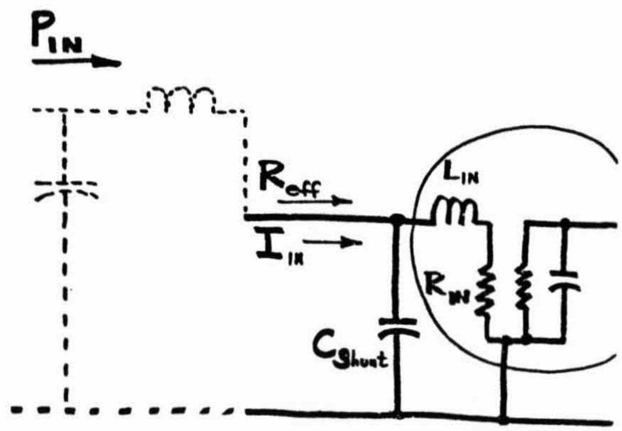


FIG. 15



$$Q_L = \frac{X_{L_{IN}}}{R_{IN}} = \frac{R_{eff}}{X_{C_s}} = \frac{f_o}{BW} = \sqrt{\frac{R_{eff}}{R_{IN}} - 1}$$

$$R_{eff} = \left(Q_L^2 + 1\right) R_{IN}$$

$$I_{IN} = \sqrt{\frac{P_{IN}}{R_{eff}}}$$

FIG. 16

CHIP CAPACITOR DIELECTRIC EFFECTS

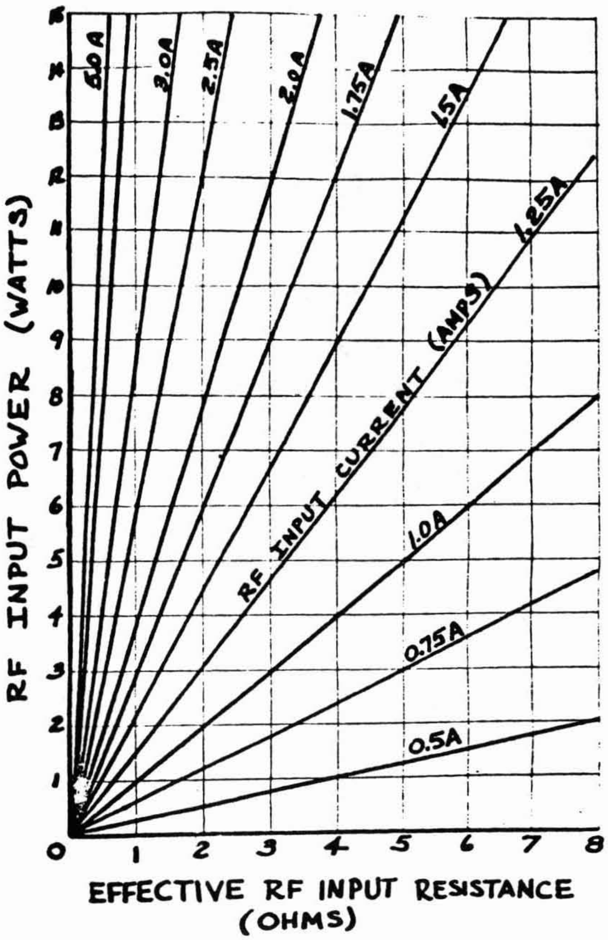
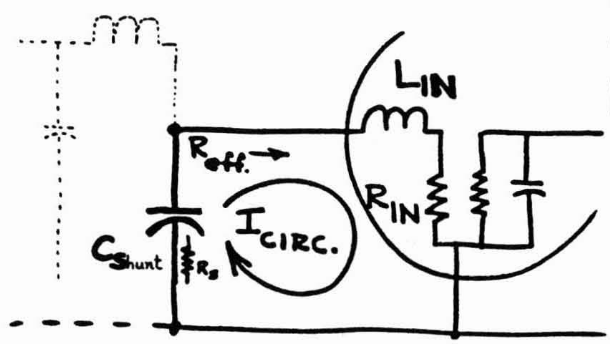


FIG. 17



$$I_{CIRCULATING} = Q_L I_{IN}$$

$$P_{diss.(in\ C_s)} = I^2_{CIRC.} R_{S(in\ C_s)}$$

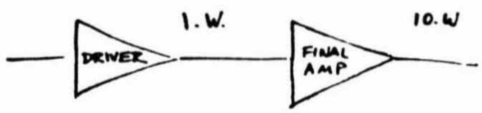
$$= I^2_{IN} Q^2_L R_S$$

FIG. 18

PRACTICAL EXAMPLE:

400. MHz 10. WATT AMPLIFIER

- RUSH BLOCK DIAGRAM FOR PROPOSAL:



- DESIGN REQUIREMENTS:

  "10. Watts output over moderate BW,
  All components as cheap as possible,
  as few transistors as possible,
  low current drain,
  minimal heat-sinking area,
  small size,
  light weight,
  short design time."

- BENCH RESULT:

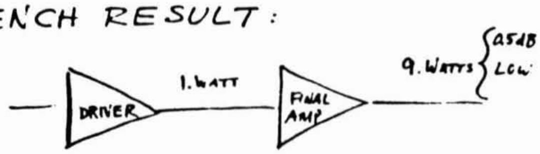


FIG. 19

PROBLEMS:

- TIME IS SHORT
- CAN'T VIOLATE DESIGN REQUIREMENTS
- MUST HAVE THAT 0.5 dB

QUICK-FIX:

. JACK UP THE DRIVER BIAS

PROBLEMS:

- η_c
- G_p
- Θ_j

FIG. 20

CHIP CAPACITOR DIELECTRIC EFFECTS

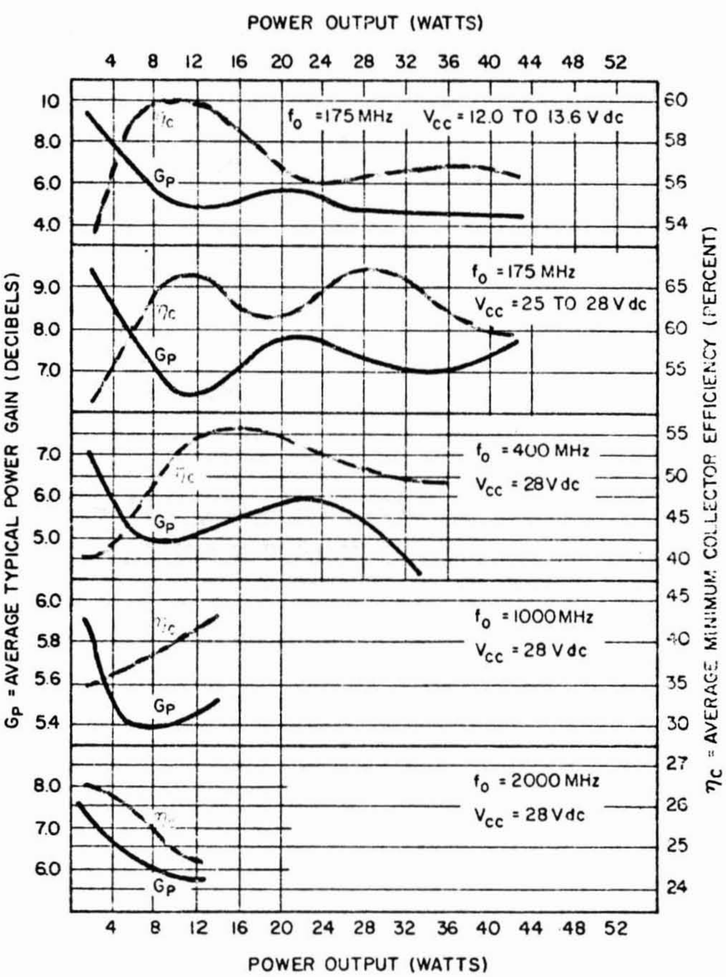


FIG. 21

2-5-23

"SIDE EFFECTS" OF QUICK-FIX:

MFR'S QUOTED η_c (MIN) = 40.%

LOST: 0.5 dB = 100. MW

POTENTIAL EFFICIENCY:

$$\frac{0.5\ dB}{40.\%} \Rightarrow 250.\ MW\ \underline{DC}$$

250 - 100 = 150. (<u>HEAT</u>)

REC'D T_j : $\underbrace{180.°C \text{ OPR.} — 200.°C \text{ MAX.}}_{\text{10.\% SAFETY MARGIN}}$

AVG. Θ_j = 24.3 (°C/W)

PROB: $\underbrace{T_j = 180. \longrightarrow T_j = 184.°C}_{\text{20.\% MARGIN LOST}}$

FIG. 22

WORRY:

FAILURE MECHANISMS:

- INSTANTANEOUS:
  - 2ND BREAKDOWN
- WEAR-OUT:
  - MASS TRANSPORT
  - ETCH PITS
  - SWELLING ALUMINUM

FIG. 23

CHIP CAPACITOR DIELECTRIC EFFECTS

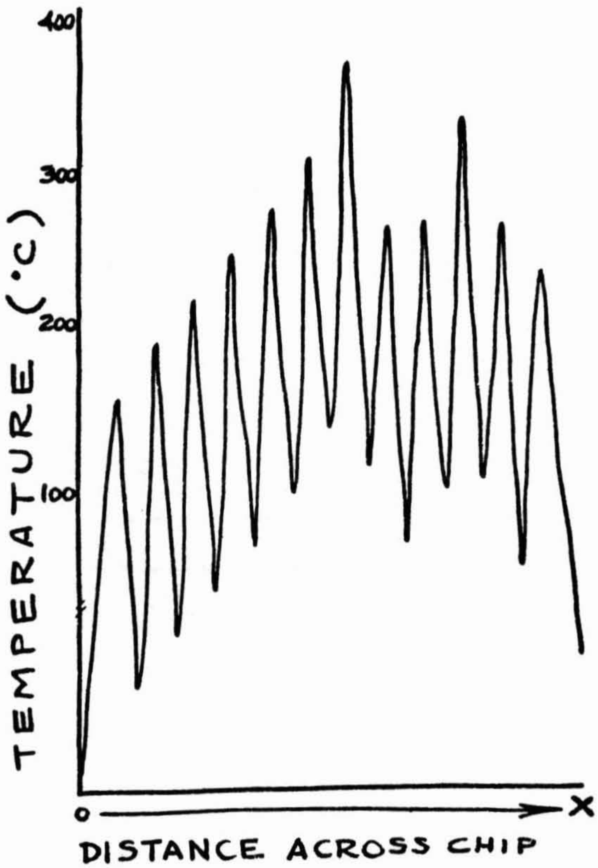


FIG. 24

2-5-24

$.001^{\underline{IN}}$ × $.001^{\underline{IN}}$ $J = 1 \times 10^{4} \left(\frac{Amp}{cm^2}\right)$

Å $J = 1 \times 10^{6} \left(\frac{Amp}{cm^2}\right)$

ELECTRON "WIND" FORCE $\propto J^2$

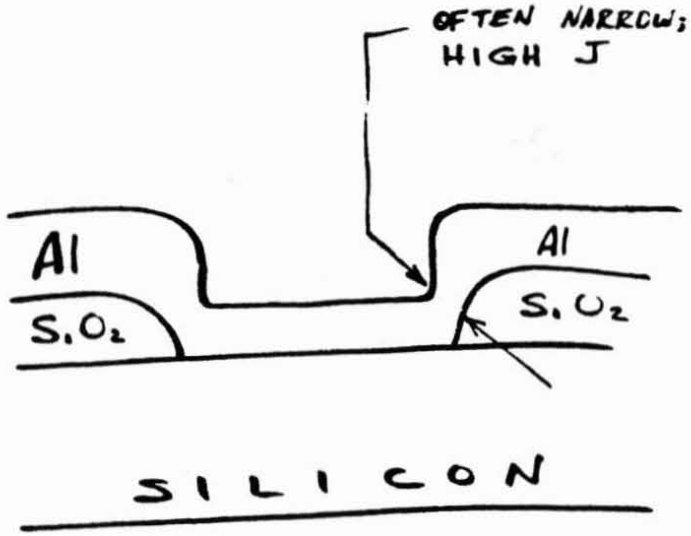


FIG. 25

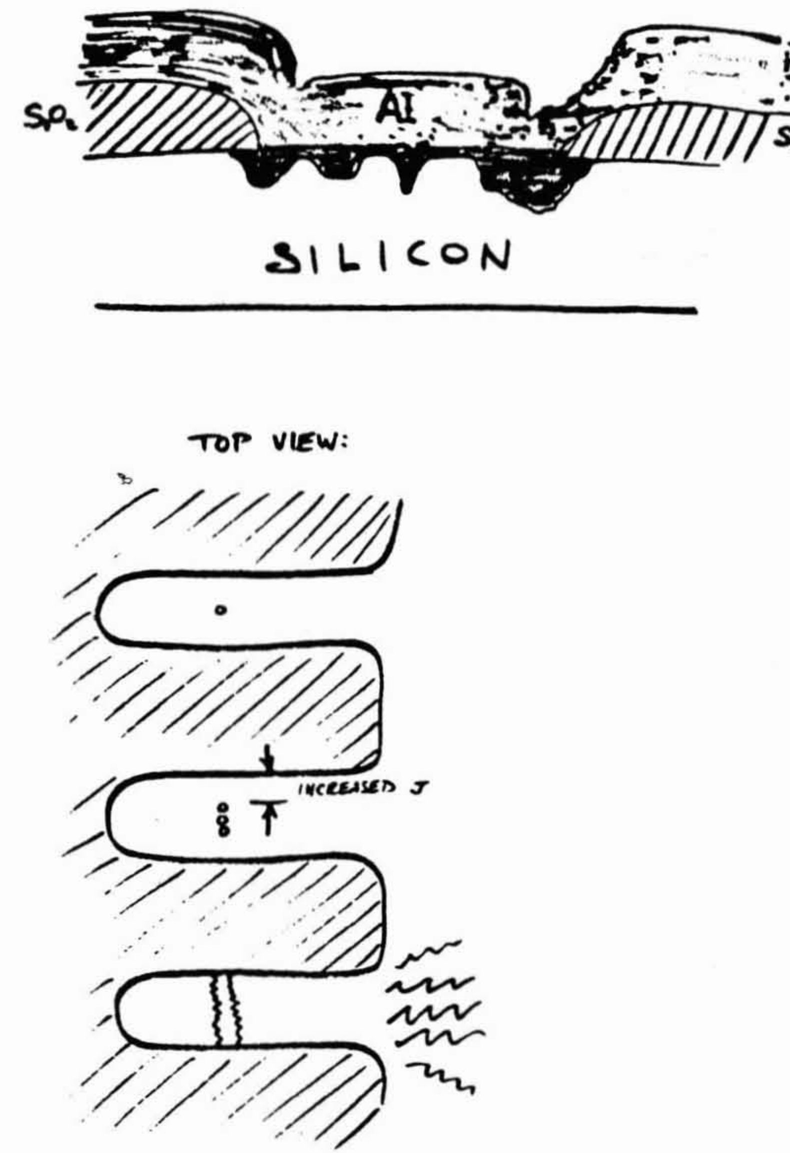


FIG. 26

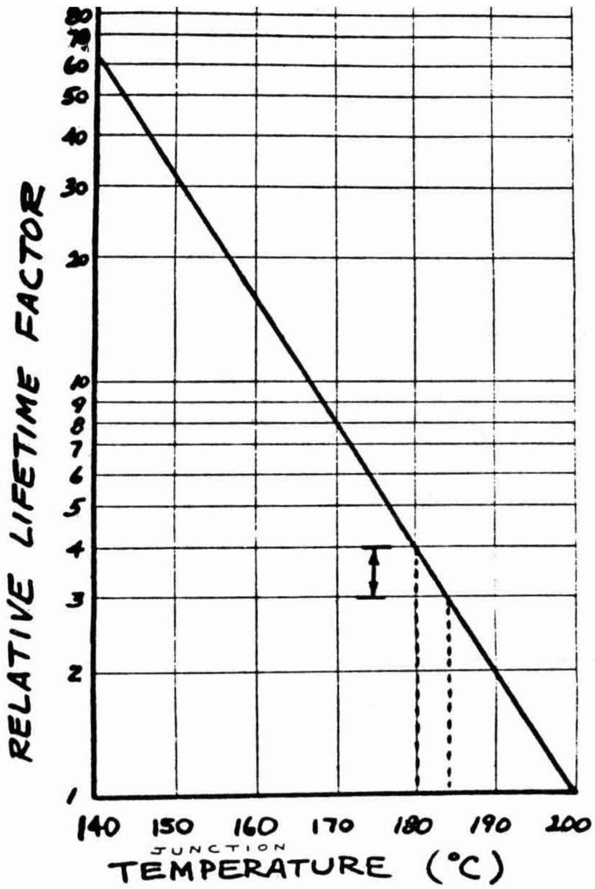


FIG. 27

CHIP CAPACITOR DIELECTRIC EFFECTS

IMPROVEMENTS REPORTED BY CUSTOMERS:

EXAMPLES (all different)

(a) POWER OUTPUT DOUBLED FROM 7. TO 14 WATTS AT 225 TO 400. MHz

(b) BANDWIDTH INCREASED BY 25.% DUE TO HIGH-END IMPROVEMENT

(c) DC. CURRENT DRAIN CUT 35.% WITH NO SACRIFICE OF R.F.

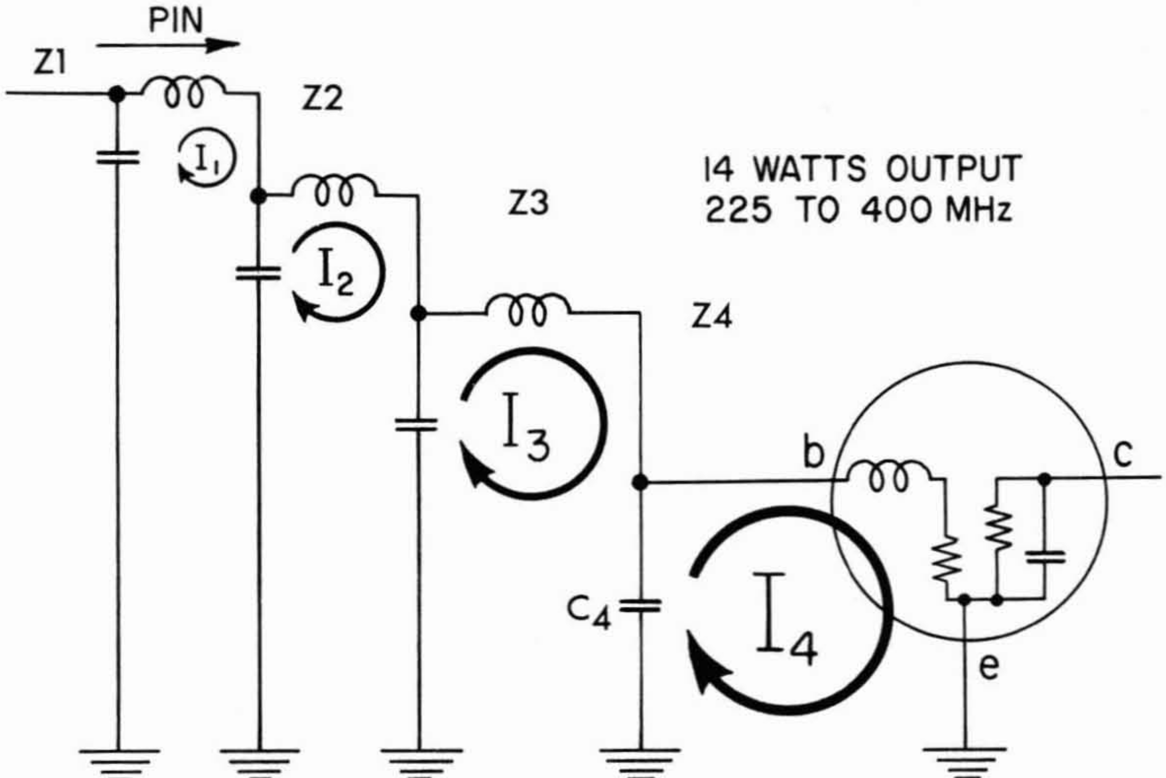
(d) POWER HANDLING QUADRUPLED TO 15. KILOWATTS AT 6.% DUTY CYCLE AT 500. MHz., WHILE REDUCING SIZE.

FIG. 28

NOTEWORTHIES FROM ATC

"TAURUS NON EST*"

DOUBLE YOUR POWER OUTPUT WITH ATC CAPACITORS



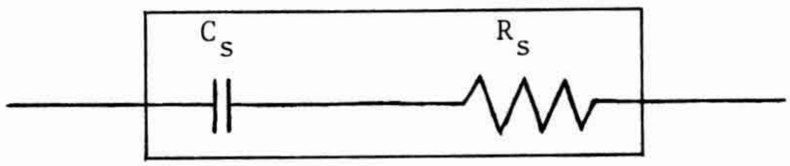
*SPANISH TRANSLATION: "EET'S NO BULL!"

IN TWO SEPARATE INSTANCES, A MAJOR INTERNATIONAL MICROWAVE SEMICONDUCTOR MANUFACTURER HAS FOUND THAT USING ATC MICROWAVE PORCELAIN CAPACITORS, <u>DOUBLES</u> THE AVAILABLE OUTPUT POWER, COMPARED TO THAT WHEN USING NPO TYPES.

Why?---Because ATC microwave porcelain capacitors have such extremely high Q that their losses (power dissipation) are nearly zero. How is this possible?---Quality Factor (Q) is the inverse of Dissipation Factor (D.F.)---both figuratively speaking and in mathematical actuality:

$$Q = \frac{1}{DF}$$

A capacitor may be schematically represented as an ideal capacitance (C_S) in series with a loss (heat) generating resistance (R_S):



The relationship between these two is defined as:

$$Q = \left(\frac{X_S}{R_S} \right)$$

where $X_S = \left(\frac{1}{2\pi f C_S} \right)$ at some specified frequency (f.)

Consequently the series resistance (R_S) is a function of Q and DF:

$$R_S = \frac{X_S}{Q}$$

or

$$R_S = (X_S)\ (D.F.)$$

What does all this mean?--Well, if a common ordinary capacitor, which has a DF (quoted as a percentage) of perhaps 0.8% at 1 KHz, is chosen for use in an amplifier at say 200 to 400 MHz, you've got miseries!

That's because the same capacitor has a loss that increases drastically with frequency, and by the time you reach 300 MHz it is at least 8.%

This means a Dissipation Factor (expressed as a straight number) of .08 at 300 MHz instead of the lesser .008 at 1 KHz.

The series resistance of the capacitor then reveals itself to be:

$$R_S = (X_S)\ (D.F) = \left(\frac{1}{2\pi f C_S} \right) \left(.08 \right)$$

EET'S NO BULL

$$= \frac{.08}{6.28 \times 3\times10^{2} \times 10^{6} \times 30 \times 10^{-12}}$$

$$= \frac{8. \times 10^{-2}}{56.6 \times 10^{-3}}$$

$$= \frac{8. \times 10}{56.6} = 1.4 \text{ Ohms}$$

"Well---what's so hot about that?", you say. Here's what:

When large amounts of RF power (P_{in}) are pumped into the "Base" of a transistor (at point "b"), the apparent resistance of the semiconductor material falls to about the 1. to 2. Ohm level.

Those little electrons are simply heating themselves to death in the R_S of the capacitor instead of serving some useful function like increasing the power output of the transistor.

This R_S is especially critical in C_4 in the diagram, because I_4 circulates both in it and the transistor, too.

And what's additionally nasty about the whole affair is that P_{in} usually starts in at some nice high impedance (Z) level like 50.Ω where not much current results, but by the time it has gotten shuffled from hand to hand thru several impedance-lowering loops of this wideband impedance-matching circuit, it will be at the 1. or 2. Ohm level of the transistor, and the current has risen to a horrendous $\sqrt{33}$ times greater than at the input. This occurs since the power level has stayed the same but the impedance level has dropped, and since "energy must be conserved" says a famous engineering lawgiver, the current rises.

Well now--That's doubly nasty, since capacitors in general have the unfortunate characteristic that heat caused by I^2R losses due to current causes DF to rise, which equates to rise in R_S, which means more heat---ad nauseum.

Now, along comes the dragon-slayer, Microwave Porcelain, with its essentially non-existent R_S (very clever people, these ATC's)and, voila!, up jumps your power output to where it should be.

A further happy sequel to this armchair adventure is that it only takes the improvement of a few such stages in a power chain and you suddenly discover you've got more power safety margin than you ever dreamed possible, and now you can look like a hero by eliminating excess stages, by saving all kinds of money, space, labor, spare parts, manufacturing time when schedules are short, current drawn from the power supply, heat, and the need for cooling plus: improving the MTBF (both yours and the circuit's).

Now the moral of this story is: cut costs where it counts. Don't be fooled by false economy. "Value Engineering" means: putting the value where it values.

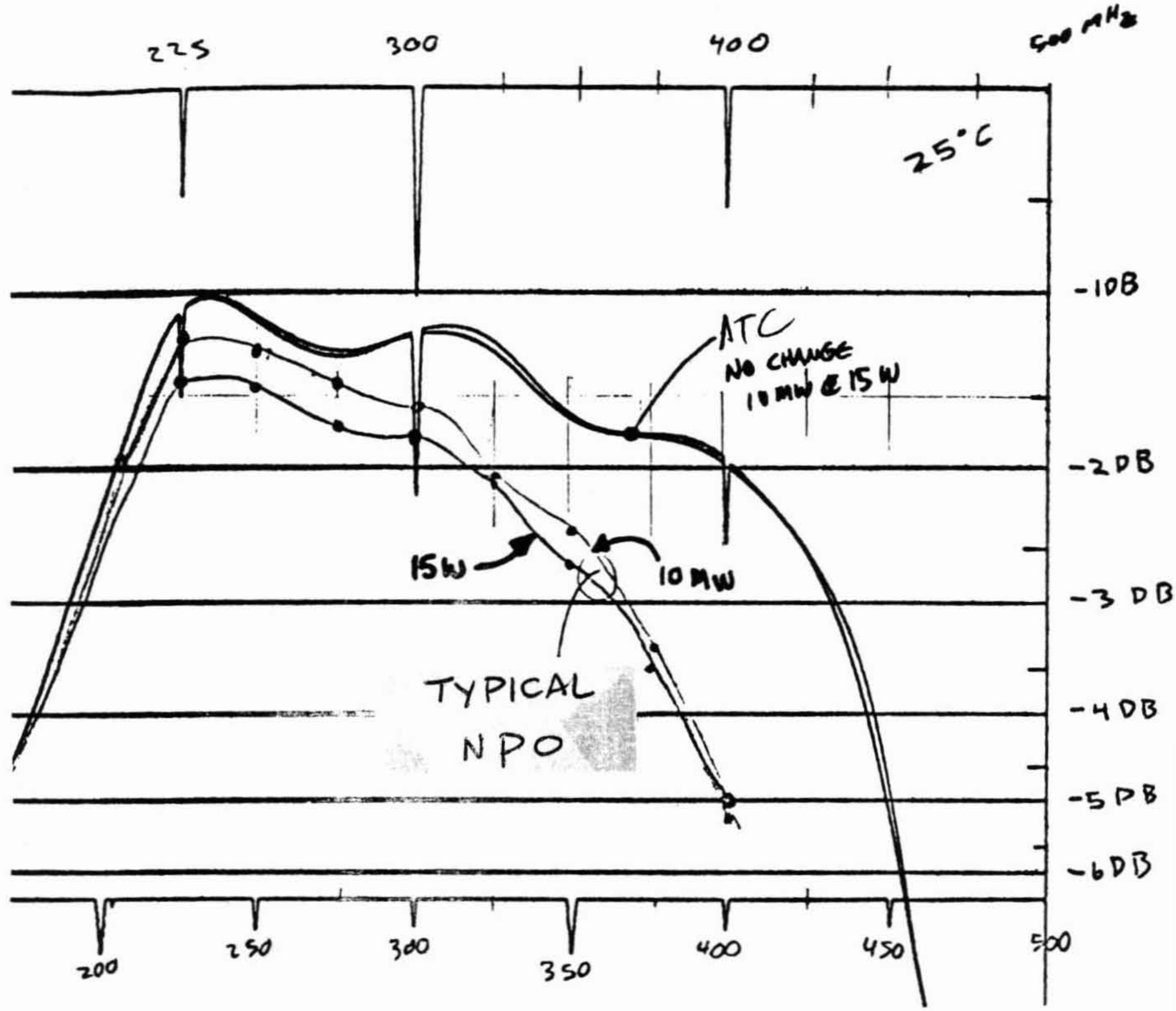
There's an old saying about quality: "you get what you pay for." For less than $1.00, you've just saved yourself $100.00 worth of headaches and money.

So when design time is short (when is it ever long?), don't mess around. There's another pithy little saying bandied much about these days": "You never have time to do it right, but you always have time to do it over." Well, don't let it happen to you.

And while we're on the subject of quality performance, it may interest you to know that the Q of an ATC microwave procelain capacitor of about 30. pF (the value used as an example earlier) at 300. MHz is 1000. or an R_s of 0.018Ω---about 100 times better than the garden-variety type.

Try that on for size.

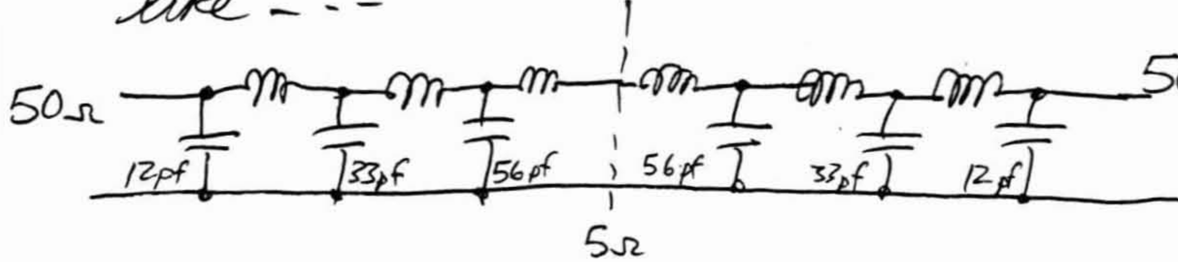
HAVE YOU WONDERED WHY YOUR COMPETITORS EQUAL YOUR PERFORMANCE WITH FEWER STAGES?



19 Oct 70

Vince,

Pardon the hand scribbled note, but I was so impressed by your ATC-100 capacitors that I couldn't wait to send you this data. Enclosed is a plot of a matching network (impedance transforming type). Two identical filters running back to back are used to take the measurements. The impedance goes from 50Ω to 5Ω and back up again. Schematically it looks like ---



As you can see six capacitors are used.

The inductors are thin film on an alumina substrate.

Two effects of changing from TYPICAL NPO capacitors to ATC-100 are evident from the graph.

1) There is a 3 db improvement at 400 mhz using ATC-100's

2) No perceivable change in ~~response~~ UNDER POWER with ATC cap's, however the other capacitors (?) seemed to change under power (SEE PLOT).

I should mention the curve will roll off at higher frequencies due to the low Q of the inductors, thus accounting for the slope in the response. Also, using the ATC capacitors, the overall circuit performs as optimized by computer. Beat that!

QUOTABLE QUOTES FROM:

"MICROWAVE FILTERS, IMPEDANCE-MATCHING NETWORKS, AND COUPLING STRUCTURES"

by G.L. Matthaei, L. Young, and E.M.T. Jones, McGraw-Hill, 1964, page 87.

"Because of their sharp cutoff, Tchebyscheff Characteristics are often preferred over other possible characteristics; however, if the reactive elements of a filter have appreciable dissipation loss the shape of the pass-band response of any type of filter will be altered as compared with the lossless case, and the effects will be particularly large in a Tchebyscheff filter."

2-9-1

ECONOMIC

(a) CONSIDERATIONS

Two major difficulties experienced by designers of high frequency transistor amplifiers are:

- the higher they go in frequency or power, the poorer becomes the gain,
- the greater the input signal level,
  - -the lower the input resistance
  - -the greater the circulating current losses
  - -the narrower the bandwidth.

Any power lost in the capacitors associated with the transistors can never be retrieved, and must be compensated for by steps which typically cost money, cut transistor lifetime, or are otherwise detrimental.

Ultra-low-loss capacitors solve these problems, and have been manufactured by American Technical Ceramics for over five years. These ATC microwave porcelain capacitors are, in fact, the lowest loss units (by at least a factor of 10) available on a production basis.

Several representative examples of customer-reported improvements in performance that resulted from changing over to ATC capacitors are:

- Double the gain and power output
- nearly 30% greater bandwidth
- 35% lower current drain
- dramatically increased power-handling capability

(b) CUTTING AMPLIFIER COSTS........... WITHOUT SACRIFICING PERFORMANCE

According to a manufacturer in the Eastern U.S., the use of ATC's microwave capacitors in place of NPO types has so increased RF amplifier efficiency in mobile radios that his required battery drain has been reduced from 20 Watts of DC to 13 Watts for the same 10 Watts of RF output power at 470 MHz...a 35% improvement; only 2/3 of the original battery requirement.

Improvements in gain and power output of 20% to 100% are also reported by other manufacturers and described in detail in ATC Application Notes* AN101 and AN104. These reiterate the fact that the use of ATC microwave capacitors allows:

- less stringent transistor gain specs, permitting the use of standard rather than selected types; or,
- a gain safety margin, or instead, a decrease in the number of stages required to attain a given RF output level;
- a decrease in DC input power or battery drain;
- the ability to experience more fully the available transistor lifetime, undegraded by excessive current and internal heating;
- a significant reduction in the <u>cost-per-Watt</u> of RF output.

All of these occur as a result of a great increase in circuit efficiency---the result of replacing power-robbing capacitors with ATC low-loss types.

To successfully accomplish significant cost savings, design work should <u>begin</u> with ATC's nearly zero-loss, true-RF capacitors. Then, <u>less expensive</u>, readily available transistors may be used at the outset, thus avoiding reworking the designs later to make up for losses. Although reduction in the number of semiconductors is possible, it is equally importantly accompanied by an even larger reduction in the number of associated components (often handmade), plus mechanical assemblies, and amplifier-chain volume and weight, as well as manufacturing <u>labor</u> charges. Add to this the reduction of time-consuming <u>fine-tuning</u> by expensive personnel in order to meet spec, and your overall savings can be 50 to 100 times the price of the ATC capacitor.

<u>In addition to the cost savings, you have a simpler, more reliable, more marketable product</u>.

(*See Sections 2-6 and 2-7)

(c) A NOTEWORTHY QUOTE FROM: "Power for Portables"*

"The two-way radio user with portable equipment has to carry his power with him. It will usually be a nickel cadmium battery, and it may comprise one-half or more of the volume and weight of his radio equipment. As advances are made in circuit construction, the battery is gradually becoming the limiting item in the quest for radios with higher power and smaller size."

*by Mr. Robert F. Chase, IEEE Transactions on Vehicular Technology, Vol. VT-19, No. 4, November, 1970, pages 248-251.

See also: Page 1 of Section 6 of Sanders Associates, Inc. Microwave Division's insert in Microwave Systems News, May/June, 1971, Vol. 1, No. 5, page 24.

See also: Page 80 of Air Force Propulsion Lab Technical Report AFAPL-TR-70-70 (Unclassified) of December, 1970, Air Force Systems Command, Wright-Patterson AFB.

(d) QUOTABLE QUOTES

——on the typical benefits derived from high quality components --- whether they be potentiometers or capacitors:

> "We made many mistakes, but we kept learning. One of our mistakes was to use bargain pots for our controls. That turned out to be very poor economy as they became scratchy with with use. Now we standardize on Allen-Bradley pots, even though they cost considerably more."**

NOTE: not only will American Technical Ceramics' low-loss capacitors *not* become scratchy, your customers will no longer get the itch to try your competitor's product "because it performs better."

**"Design Interface", EDN/EEE, May 15, 1971, page 58.

(e) QUOTES, ABSTRACTS, AND COMMENTS ON

"MICROWAVE WARFARE.... WITH EXPENDABLE JAMMERS"

"The ingredients for success are accurately tunable microwave circuits....and high energy-density battery sources _____ all with a low price tag."

"Costs are only one problem....Getting enough microwave power and good efficiencies out of a very restrictive volume is also very challenging."

"Above S-Band, Gunn and Impatt sources are considered as fundamental microwave sources as an alternative to multipliers. The major disadvantage of bulk devices, however, is their lower efficiency."

"Another critical problem facing the development of expendables is that of finding battery sources which can offer suitable voltages for reasonable periods of time."

((Note: This is echoed by an Air Force Report which discusses the problem that, because of the increasing demands for DC current and higher power output, small batteries are becoming a significant portion of the physical size of ECM and other small-system packages. An example of present-day limits on battery size, weight, and discharge-rate is: 6 cubic inches, weighing one-half pound, providing only 1 Ampere at 28 volts DC for 10 minutes. Cost: as much as an expensive microwave transistor.))

"Another major problem in the development of expendables is antenna design. Since the 'effective radiated power or ERP' is already constrained by existing battery sources and the available output from present day semiconductors, it is desirable to couple in as much jamming power as possible."

(Since DC-to-RF conversion efficiency has such a large influence on Effective Radiated Power, it is imperative that the coupling and tuning capacitors of expendable jammers have the lowest possible RF loss, which means the highest Q, <u>at</u> the frequency <u>and</u> RF current level of intended operation.)

NOTE: ATC capacitors can handle extremely high current-densities while <u>simultaneously</u> providing significant fractions of a dB (and in some cases, up to 3 dB) more RF power output at UHF and above than any other capacitor made. This can have a very favorable impact on the jammer's cost-effectiveness.

REF: "Microwave Warfare (Part 2): Spoofing Hostile Radars with Expendable Jammers" by R.T. Davis, MicroWaves Magazine, October, 1971, pages 34 - 37.

Page 80 of Air Force Propulsion Lab, Technical Report AFAPL-TR-70-77 (unclassified) of Dec, 1970, Air Force Systems Command, Wright-Patterson AFB.

(f)

<u>Quotes and extracts from: "The Hand-Helds:'communicating to the man'"</u> Electronic Packaging and Production, June, 1972, page 45. (See RF Packaging: two approaches")

"Hand-held 2-way FM radios are products distinguishable by the restrictions imposed on their design. The approach is mainly that of obtaining an optimum efficiency from packages severely restricted in volume and weight.

Their duty cycle is described (EIA) as 10-10-80, that is, 10% transmit, 10% receive and 80% standby at full rated RF power output (maximum about 5W, minimum about 1W). This translates into 8-12 hr. of efficient use, after which the radio's battery must be recharged. It is, in fact, this limiting factor of the battery that prevails: Any breakthrough in radio package design really awaits some progress in battery development....or, most importantly,....an improvement in time-power efficiency" ((in the semiconductor RF power output stages.))

QUOTES AND EXTRACTS FROM:

(g) A 140. WATT, 175. MHz AMPLIFIER WHERE LOW-LOSS CAPACITORS ARE A MUST

(by R. Cushman), EDN/EEE, November 15, 1971, page 34

"This land-mobile communication amplifier by Communications Transistor Corp., San Carlos, Calif., must use low-loss, low-parasitic inductance capacitors..." (since) "...the circulating currents are several Amperes..."

"Losses caused by an ineffective first match may cause 3.dB loss in gain as well as saturated power output. Gain is too hard to come by at 140.W and 175.MHz to throw it away in a passive component."

CTC engineers have found that ATC low-loss porcelain capacitors "have better RF dissipation (higher Q_u's) and lower parasitic inductance than the more common barium titanate NPO types often used.

"Though these porcelain capacitors are about three times as expensive as the NPO types, their cost is still small compared to the expensive transistors needed to produce large powers at these frequencies."

(h) DO YOUR PRESENT CAPACITORS GIVE YOU DOUBLE YOUR PREVIOUS OUTPUT?

QUOTE FROM: Motorola Semiconductor Products Inc. Application Note #AN-548 on a 25. Watt, 450-512 MHz, 12.5 VDC, RF Power Amplifier using 2N5945, 2N5946, and 2N6136 Transistors:

"The use of the" (ATC) "porcelain dielectric chip capacitors for the series elements in the interstage networks was found to provide an additional 2.5 to 3.0 dB of gain over that obtained with compression trimmers, as well as reducing the number of tuning adjustments necessary." [Note: "3 dB of gain" = <u>double</u>]

(i) "A LITTLE NEGLECT MAY BREED GREAT MISCHIEF"

[--Poor Richard's Almanac by Benjamin Franklin (1758)]

By Vincent F. Perna
American Technical Ceramics
Vice President
Microwave Engineering

Through long experience it has been found that great caution is advisable when considering the use of a cheaper component, especially in Military hardware. That people nevertheless seem inclined to "cut corners" is nothing new, having been commented upon even by writers in the Middle Ages.

Apparently some embattled monarch got a bitter pre-taste of the need for MIL-Spec's when he encountered even the blacksmith's reluctance to put full value into his horse-shoe nails. This is evidenced by a pithy lament that immortalizes some unknown farrier's negative contribution to the war effort by trimming a little on the quality of his pig-iron (probably motivated by the ancient equivalent of the adage: "a penny saved is a penny earned.") This commentary on the True Cost of Ownership of components of not-quite-optimum quality mourns:

"For want of a nail, the shoe was lost,....
"For want of a horse, the rider was lost,....
"For want of a battle, the kingdom was lost,
"-----And all for want of a horse-shoe nail."

In more recent times, to avoid such disasters, MIL-Spec's eventually did come into being, but in some cases seem to have gotten out of hand, leading to the tongue-in-cheek story that an elephant is a mouse built to MIL-Specs.

Pachyderms are, of course, fine in their place, but the average non-Military user of high frequency electronic equipment is generally willing to accept an animal-size somewhere in between. Unfortunately, when his less ponderous thin-skinned brother goes into commercial applications, the age-old problem of unnecessary breakdowns re-surfaces.

One difficulty with the slide-it-on-by type of manufacturing philosophy is, however, that the repairman may be getting paid at a significantly higher rate than the man who designed the equipment-----which is a peculiar kind of economics. (One radio company reportedly pays out a half-a-million dollars *annually* in in-warrantee repair claims.) (After that, the customer pays.)

Furthermore, as recent Government response to the aspirations of the rising generation of customers has demonstrated, stress is now being laid on how the product affects the end-user (i.e. the customer bites back.)

Since a failure puts a severe dent (or worse) in the customer's plans----about which failure he no doubt speaks out freely----it seems monetarily appropriate to recall the slogan "A Satisfied Customer is the Best Form of Advertizing." The

NEGLECT BREEDS MISCHIEF

converse is also true, and can be costly to overcome when a brand name gets itself in general bad odor.

The solution? Where RF capacitors are concerned, ATC 100 High Q porcelain units earn their keep by providing your transmitters with more gain, wider bandwidth, more power output and greater reliability with no extra effort, plus giving longer lifetimes to your semiconductors since now you don't have to push them so hard to overcome circuit losses. As a result, you can also often eliminate several other components and their associated labor costs.

To shorten your time from design to production, we provide in addition to product quality, thoroughly detailed technical characterization of our components. This can visibly reduce your initial labor costs and mean avoidance later of bottle-necks and crash-re-work programs caused by component and design "bugs" that surface after your equipment starts rolling down the production lines.

In keeping with our tradition of quality, we have continually worked to maintain excellence of service plus fastest delivery. We strive to make your job easier in all ways.

Your local ATC representative stands ready to show you how you can profit from using our low-loss procelain capacitors.

MICROWAVE AMPLIFIER DESIGN*

HOW TO AVOID GAIN AND POWER LOSS THROUGH PROPER CAPACITOR SELECTION

Typical microwave power amplifiers suffer hidden gain and power loss in components outside the transistor. Particularly detrimental are the losses incurred if the engineer uses capacitors normally suited only for audio frequencies. Losses may amount to several dB if care is not taken in the choice of dielectric Q.

Most manufacturers of low-frequency ceramic capacitors quote dielectric characteristics in terms of Dissipation Factor (DF) at 1 kc. This data has limited usefulness at microwave frequencies. To benefit from a knowledge of Q and DF, calculation must be made of the equivalent loss in terms of RF series resistance in the capacitor, and it must be from data at the anticipated frequency of operation. Most ceramic capacitors do not have published data above 100 MHz. If operation at microwave frequencies is desired, an uncertain extrapolation must be made. For example, the DF of ordinary NPO material is of the order of .0003 in the audio frequency range. At L-Band, however, this has increased by nearly 4 orders of magnitude, resulting in a DF of about 0.9 according to one manufacturer.

Performance Comparison

To make a comparison between the performance of a microwave porcelain capacitor dielectric and an audio frequency ceramic dielectric, we must employ the relations: $Q = (1/DF) = 1/\tan \delta$).

The highest attainable Q in a capacitor is necessary in order to realize the fullest gain and power from the transistor. To determine the losses, we must examine the function of the capacitor in the circuit. Typically, it is used in impedance matching, and may be either in series or in shunt with the transistor input. The effective input resistance (R_{eff}) of most broadband power transistor networks is of the order of 2Ω or less, that of very high power amplifiers being of the order of 0.25Ω. Any series resistance (R_s) in a capacitor used for coupling or matching will act as a power divider in combination with the input resistance of the transistor.

When matching down from a high impedance to a low, the power stays the same; therefore, the current level must rise. A typical current multiplication ratio is greater than $\sqrt{33}$. At a 2 Watt input R_{in} level, the RF current of the system 50Ω impedance level, transformed down to that of the transistor input, will result in over one Amp of input current. The resulting I^2R_s losses thus become extremely severe.

RF Capacitor Resistance

The RF resistance of a capacitor may be determined from $R_s = (X_c DF) = (X_c/Q_{cap})$, where Q_{cap} is the component Q, and X_c is the series reactance of the capacitor at the frequency of interest. For example, typical capacitor values of 30 pF might be encountered in such a situation, and if the dielectric used has a DF of 0.8 at, say, 300 MHz, the resulting equivalent series resistance in the capacitor will be about 1.4Ω. This equivalent resistance in combination with the 1 to 2 Ohms of the input resistance of the

*(An expanded version of an article by Vincent Perna, Vice President, Microwave Engineering, printed in Microwave Systems News, Sept./Oct., 1970)

*(See: Sections 2-6 and 2-7.)

transistor network results in several dB of dissipation loss, cutting both gain and power output. In addition, this RF resistance increases with temperature. The detrimental effect this has on drive-level sensitive transistors is dramatic, in some cases preventing turn-on when RF input or DC level drops, or when temperature rises.

The RF current resulting from the large impedance transformation from the 50Ω level down to that of R_{eff} may be calculated from $P_{in} = I_{in}^2 R_{eff}$ and $I_{in} = \sqrt{P_{in}/R_{eff}}$. For a 4 Watt input and an R_{eff} of 1.5Ω, this results in over 1.6 Amps of input current. (This is then increased to I_{circ} by the loaded Q of the network.) Microwave capacitors with a Q of 1000 can result in a capacitor R_s of only .018Ω at 300 MHz whereas low-frequency capacitors with a DF of .04 (a Q of 25) result in an R_s of 0.7Ω at that frequency. Note that at UHF, the low-frequency ceramic-dielectric has 40 times the series AC resistance of that of the microwave porcelain.

For the DF's mentioned in this case, the power wasted in the low-frequency dielectric would be over 1 Watt, whereas that dissipated in the microwave dielectric would be only 25 milliwatts.

The resultant heating in low-frequency capacitors is all the more intense when we consider that it is not spread throughout the volume of something so massive as a 2 Watt carbon composition resistor, but is rapidly building up the temperature of perhaps a small disc-ceramic capacitor 1/4 inch in diameter and 70 mils thick.

The total input power to the tuning circuit (assuming no loss in the inductor), therefore, is that needed to turn the transistor on, plus that needed to allow for dissipation in the capacitor; thus, from Figure 2:

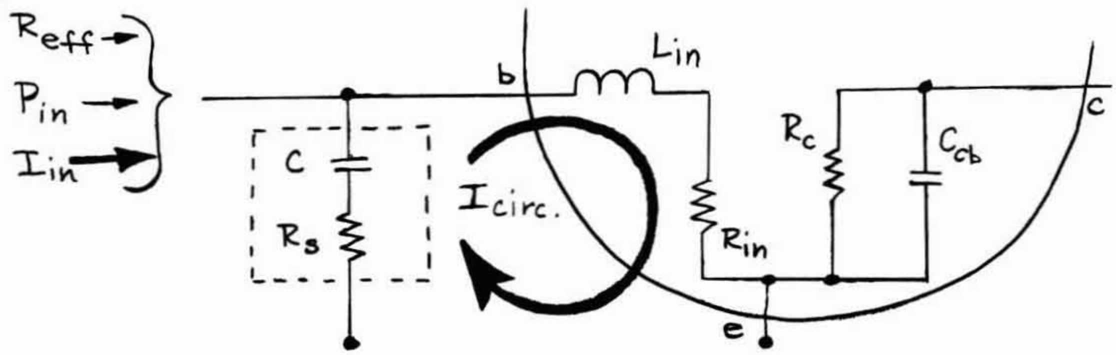


Figure 2.

$$P_{in}\ (\text{req'd}) = \left[P_{transistor} \Big/ \left(\frac{R_{in}}{R_s + R_{in}} \right) \right]$$

Consequently, (again assuming no loss in the inductor) the dissipation loss incurred due to the capacitor in a matching-section at the input to an RF transistor would be given by:

$$\text{Loss (dB)} = 10_{\log 10} \left[(R_s + R_{in}) / (R_{in}) \right]$$

Thus, for proper amplifier operation, consideration must be given (early in the block diagram stage) to the characteristics of the capacitors anticipated for use, in order that the actual gain later attained by the circuit may be nearly that of the capability of the transistor. This approach will increase the circuit DF-to-RF conversion efficiency, increase the gain and power output, and increase the transistor MTBF, since there will be less heating in the transistor for the same output power. In addition, system cost will be reduced.

INSERTION LOSS, CAPACITOR Q, AND THE S-PARAMETERS

VINCENT F. PERNA
Vice-President
Microwave Engineering
Huntington Station, N.Y.
516-271-9600

Although most microwave engineers are quite familiar with the concept of "Insertion Loss" (I.L.), its actual relationship to the various Scattering Parameters is less clear, as is the non-independence of reflected and transmitted powers.

For example, glancing at the I.L. column of the tabulated test data of Figure 1, it might appear that a 10 pf capacitor at 100 MHz would have a poor Q (= Q_{cap})

FIG. 1-TEST DATA ON 10 pf ATC 100-A-100-J-C-50

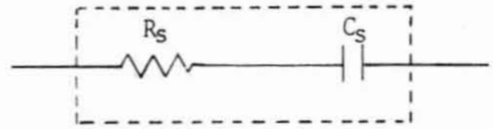
FREQ. (MHz)	FORWARD VSWR	FORWARD GAIN	S-MAGN AND ANGLES 11		S-MAGN AND ANGLES 21	
100	11.513	-5.24	.840	-32	.547	55
150	5.960	-2.61	.713	-43	.740	44
200	3.962	-2.16	.597	-51	.780	36
250	3.039	-1.16	.504	-56	.875	27
300	2.547	-1.07	.436	-59	.884	25
350	2.244	-0.89	.383	-62	.903	20
400	2.037	-0.66	.342	-64	.927	19
450	1.883	-0.72	.306	-64	.921	16
500	1.772	-0.51	.278	-65	.943	16
550	1.685	-0.45	.255	-66	.949	14
600	1.617	-0.28	.236	-67	.968	13
650	1.554	-0.24	.217	-67	.972	11
700	1.501	-0.23	.200	-68	.974	10
750	1.458	-0.22	.186	-67	.975	9
800	1.424	-0.27	.175	-67	.969	9
850	1.397	-0.21	.165	-68	.976	8
900	1.372	-0.23	.157	-68	.974	7
950	1.355	-0.18	.151	-69	.980	7
1000	1.338	-0.18	.144	-70	.979	6

(= Q_u) - but one that, strangely, improves rapidly with increasing frequency. Considering the definition of Q:

$$Q_u = \left(\frac{X_{CS}}{R_S}\right) = \left(\frac{1}{2\pi f C_S R_S}\right)$$

Reprint from MICROWAVE SYSTEMS NEWS
March/ April 1971

for the capacitor model:



it becomes clear that, for fixed R_S and C_S, Q decreases linearly as frequency increases. Consequently, one is forced to seek clarification of the extraordinary rise in I.L. (seemingly very low Q_{cap}) evidenced at low frequencies. (It is the nature of dielectrics to exhibit increasing losses with increasing frequency, and possibly falling capacitance with increasing frequency. At low frequencies, these shifts tend to cancel each other out and do not provide a suitable explanation of radical changes in I.L.) One answer comes from the VSWR column of Figure 1. If we plot these several functions versus frequency, what appears to be dissipated-power loss is actually due to the high capacitive reactance choking off the signal flow, due to the high reflection of power being induced. (See Figure 2) The transmission loss curve shown is the insertion loss minus VSWR effects, and is that which would prevail if no power were lost due to reflection.

For a test fixture consisting of a simple capacitive DC-block on a transmission line, the load seen by the generator becomes $Z = (R_S + X_C + 50\,\Omega)$, and as such can have a high VSWR given by:

$$\text{VSWR} = \left(\frac{1 + |\Gamma|}{1 - |\Gamma|}\right), \text{ where } \Gamma = \left(\frac{Z - 50\,\Omega}{Z + 50\,\Omega}\right)$$

Thus, the larger the Z due to either X_C or R_S, the larger the VSWR and therefore the larger is the reflected power.

Since I.L. is a combination of reflected and dissipated power, if either portion is high, the composite will appear high.

The Microwave Engineer's Handbook (by Horizon House) has a set of tables on Γ, reflected power, and transmission loss due to reflection ... all three versus VSWR.

Similiar information is derived from:

$$\text{VSWR} = \left(\frac{1 + |S_{11}|}{1 - |S_{11}|}\right)$$

where the load reference is $50\,\Omega$, and S_{11} equatable to the voltage reflection coefficient. Likewise, pertinent information on the power reflected by an impedance mismatch is obtainable from:

$$P_{mismatch}\ (\text{db}) = 10\log_{10}\left(1 - |\Gamma|^2\right)$$

and similarly:

$$\%\ \text{Power Loss} = \left(1 - |S_{11}|^2\right) \times 100\%$$

Power lost through reflection does not enter the capacitor and be transmitted. It thus is not involved in power loss due to dissipation or radiation in the capacitor. Keeping these points in mind, examing the test data will give a feeling for these relationships.

INSERTION LOSS, CAPACITOR Q, AND THE S-PARAMETERS

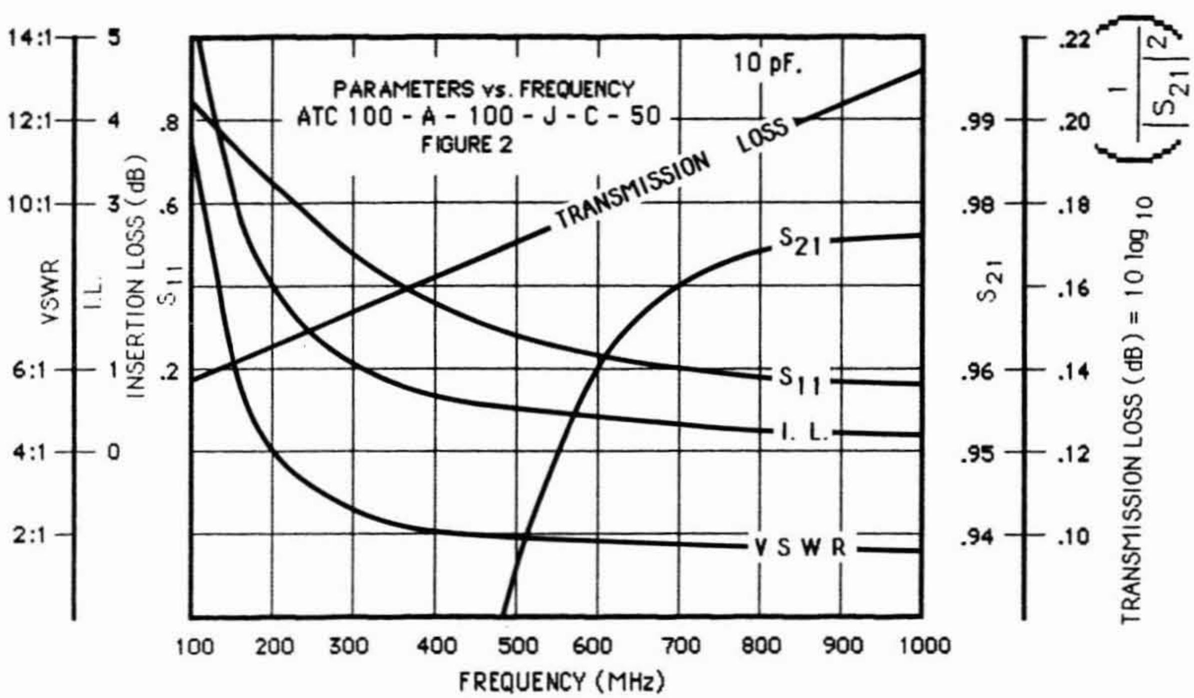


FIGURE 2

Specifically, let us examine the 100 pf capacitor (Part # ATC 100-A-101-J-C-50) listed on the front of AN102 and reproduced for clarity in Figure 3.

FIG. 3.-ATC 100-A-101-J-C-50

FREQ. (MHz)	FORWARD VSWR	FORWARD GAIN	S-MAGN AND ANGLES S_{11}	S-MAGN AND ANGLES S_{21}
100	1.369	-.11	.155	.987
150	1.234	-.01	.105	1.001
200	1.173	-.03	.080	.997
250	1.138	-.02	.065	.998
300	1.116	-.01	.055	.999
350	1.100	-.02	.048	1.002
400	1.090	-.01	.043	.998
450	1.082	-.02	.039	.998
500	1.077	-.02	.037	1.002
550	1.072	-.01	.035	.999
600	1.068	-.02	.033	.998
650	1.064	-.02	.031	.998
700	1.060	-.01	.029	.999
750	1.057	-.02	.028	.998
800	1.054	-.02	.026	.997
850	1.050	-.02	.024	.998
900	1.047	-.03	.023	.996
950	1.043	-.02	.021	.997
1000	1.039	-.01	.019	.999

At 1000 MHz this capacitor exhibits a "negative gain" (=I.L.) of .01 dB. From our discussion of Q earlier, we would therefore expect that at 100 MHz the I.L. would be so low (the Q so high) as to require that we go beyond two decimal places in the tabulation of I.L. in order to see any evidence of loss. Additionally, the lower one goes in frequency, the lower a portion of wavelength the dimensions of the capacitor would appear to be, and thus the lower the likelihood of the losses ascribable to radiation. We can therefore safely assume that the indicated I.L. is due almost exclusively to reflected power loss. The same principle applies in the case of the 10. pf capacitor in Figure 1.

Although the basic I.L. is higher at 1000 Mhz, the dissipative losses can be assumed to be a negligible portion of the I.L. of 5.24 dB at the 10 times lower frequency of 100 MHz. The remaining loss is due to reflection.

Consequently, when a quick evaluation of relative capacitor Q's must be rendered on the basis of transmission line DC-block data, it is advisable to examine the low VSWR region of the frequency spectrum in order that the dissipative portion of the Insertion Loss not be masked by reflected power effects.

A NOTEWORTHY QUOTE FROM "A NEW LOOK AT MICROWAVE SILICON TECHNOLOGY": *

"Excessive heat is by far the main cause of transistor degradation and failure. For semiconductor devices it is the usual practice to base the maximum power rating as the theoretical allowable temperature the junction can attain, 250°C for silicon. Beyond that temperature, conduction becomes intrinsic and the junction loses its rectification properties. A conservative junction-temperature rating is 200°C for silicon devices."

"The use of extremely fine surface geometry can be expected to maximize the cutoff frequency (f_t). However, the large current densities consequent upon the fine structure provide increased problems...."

"Higher gain at the expense of lower efficiency is common at lower power levels."

*by E.J. Rice, R.D. Gromer, K.M. Finn, B.J. Tilley, and W.E. Schaub, The Microwave Journal, February, 1969, pp. 80-86

QUOTES AND ABSTRACTS FROM: "RF POWER TRANSISTOR METALIZATION FAILURE"

"RF power transistors employing aluminum metallization can degrade and fail when the aluminum carries high-current densities at elevated temperatures".

Under these conditions, the sheet resistance of the aluminum conductors increases and the beta falls off, reducing the device efficiency.

Especially prone to failure are: "small emitter area devices containing shallow junctions," and, "high frequency devices and particularly high-frequency power devices."

"Due to the solid solubility and relatively rapid diffusion rate of silicon in aluminum, the latter can act as an etchant for silicon".

Etch pits grow into the silicon under normal device operating conditions, resulting in device "wearout".

Silicon ions are swept away (mass transport) from the silicon-aluminum interface and carried down the aluminum conductor by "electron wind" forces.

"As the aluminum near the silicon interface is depleted of dissolved silicon, more silicon from the substrate is able to dissolve into the aluminum. The process is a continuous one, resulting eventually in silicon etch pits filled with aluminum which may penetrate a neighboring junction, resulting in an electrical short".

The silicon swept out of solution deposits where the electrical path widens and the current density drops. **

For long RF transistor lifetime, "the current density in the metallization and the metallization temperature must be maintained below the levels which cause rapid electromigration in aluminum ($J \leq 1 \times 10^5$ A/cm^2 and $T \leq 125°$C)."

---

* by J.R. Black, IEEE Trans. on Electron Devices, Vol. ED-17, No. 9, September, 1970, pages 800-803.

** (Note: these deposits build up surface roughness and resistance, rapidly degrading transistor performance.)

Watch those losses in low-power amplifiers.

Otherwise you may be rudely surprised to find that your transistors aren't lasting as long as they should.

All engineers know that efficiency is very important in high-power rf amplifier design. What is often not appreciated is that unexpected losses can have catastrophic consequences in low and medium-power circuits as well.

The catastrophe comes about because economic considerations usually dictate that an amplifier should contain the smallest possible number of transistors, all pushed very close to their operating limits. Then, if a small unexpected loss is encountered, transistor failure—either immediate or slightly delayed—is almost sure to follow because of the "quick fix" measures that are commonly employed to make up for the loss.

The most tempting way to compensate for the loss is to push a little more dc bias through the transistor to pick up the needed additional gain. And here is the often-unexpected rub: *Low-power, high-gain devices tend to be significantly less efficient than their high-power, low-gain counterparts* (Fig. 1). Hence, getting a little more power out of one of them may involve dissipating quite a bit more heat. And if the junction temperature is already operating near its safe limit, disaster is just a short step away.

Direct heating is a problem, too

Lossy interstage components can create thermal problems in yet another way: They can simply heat up the transistors directly. For wide bandwidths, impedance-transforming networks of the type shown in Fig. 2 are often used.

Although each capacitor causes some loss, C_1 is the most critical, since it must be placed directly at the transistor case, where the metallic lead enters the device. This ensures that the input inductance will not cause too great an impedance step-up, which would increase the amplifier's loaded-Q. Since P_{in}, the input power, remains fixed despite the drop in impedance level, circulating current I_1 may be on the order of several amperes.

Significant amounts of heat can be generated in C_1 if the capacitor's dissipation factor is poor (in some types, it's enough to melt the solder). For example, if its equivalent rf series resistance were only 0.1 Ω, an input of only a few watts would cause about 0.25 W of dissipation in the capacitor.

Because of the intimate connection between this capacitor and the transistor, they mutually degrade each other's operation—the capacitor adding to the transistor's heat burden, and the transistor's heat degrading the capacitor's Q (usually exponentially with temperature). Since the transistor is usually operating at its maximum safe power level (T_j typically hovering just at or above 180°C), any heat input from sources outside the transistor will eat into this safety margin, making the junction more likely to develop hot spots that will lead to degradation and eventual failure.

0.5 dB may be bigger than you think

To see just how serious these seemingly simple considerations really are, let us assume that we wish to design an inexpensive amplifier capable of producing 5 W at 400 MHz from a 28-V dc supply. According to Fig. 1, the average output transistor will have about 7 dB of gain. Therefore the driver stage will have to deliver about 1 W of rf power.

Now, if we are unaware of the existence of extremely low-loss capacitors for the interstage network, we may be rudely surprised to discover that there is, say, a 0.5-dB transmission loss between the driver and the output stage.

Two quick fixes come immediately to mind:

1. Pull out the driver and replace it with a higher gain—but more expensive—device.

2. Bias the driver closer toward saturation to pick up the needed gain.

Since higher-gain devices have lower collector efficiencies (Fig. 1) and cost more, and we already have a working device in the circuit, we choose to push more dc through the one we have.

Vincent F. Perna Jr., Vice President, Microwave Engineering, American Technical Ceramics, 1 Norden Lane, Huntington Station, N. Y. 11746.

Reprinted from Electronic Design 10, May 13, 1971

Keep cool: Heat can kill transistors in two ways

Operating a transistor at too high a temperature can lead to device failure in these ways:

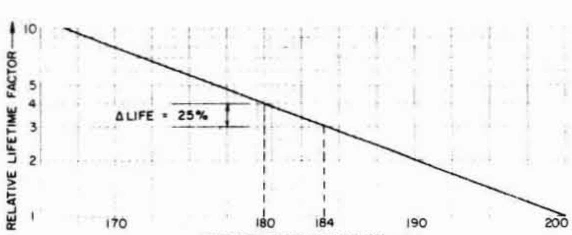
- Secondary breakdown—a catastrophic failure mechanism in which the collector and emitter get shorted together.
- The mass transport phenomenon—a more subtle occurrence in which the transistor's lifetime is greatly shortened by the migration of its metallization under the influence of both high-temperature and high-current density.

The effect of junction temperature on transistor lifetime, through the mass-transport mechanism, is profound. Going from a junction temperature of 180° to 190°C, cuts the device life in half. And going to 200°C cuts it down to a quarter of the 180°C value. At this rate even the seemingly small temperature rise of 4°C (cited in the example in the text) will reduce the transistor's lifetime to 75% of what it would have been at 180°C (see illustration).

Secondary breakdown is sudden death

Secondary breakdown is the result of localized current crowding that raises temperatures enough to melt the semiconductor material. As the material melts, its resistivity is drastically reduced, and the collector-to-emitter junction takes on the electrical characteristics of a metallic conductor. This typically occurs between 200 and 300°C, depending upon the material.

Although manufacturers usually specify 200°C as the maximum safe junction temperature, $T_{j(max)}$, it should be borne in mind that the temperature at a localized hot spot may be well above the average. If this occurs, the hot spot can trigger secondary breakdown at that point.



Consequently most manufacturers recommend keeping T_j well below 180°C for reliable long-term operation.

Secondary breakdown is energy dependent—that is, it is a function of time, voltage and current. To fight it, therefore, some devices are provided with curves showing safe operating conditions. When these are not available—as is often the case with microwave transistors—the engineer must be exceedingly careful.

Mass-transport phenomenon works 2 ways

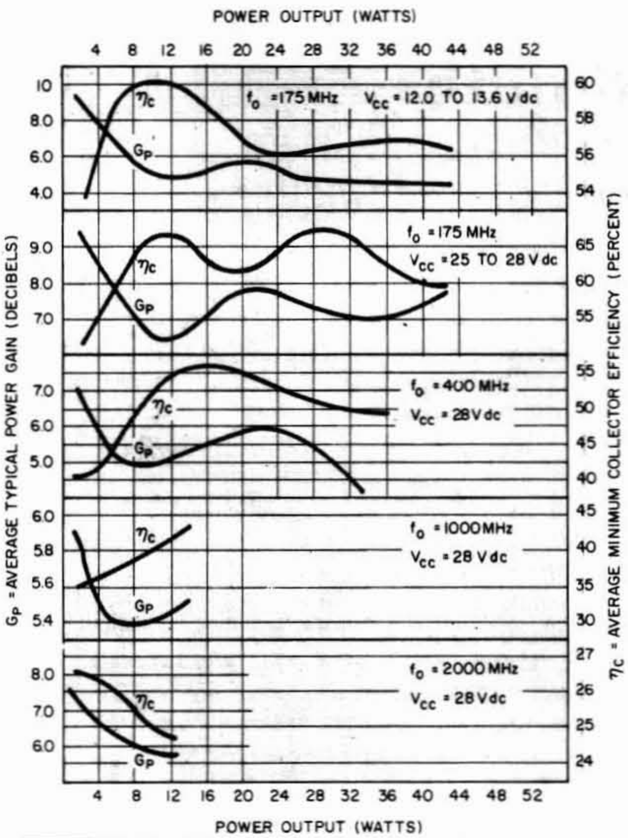
Silicon devices with aluminum metallization exhibit two distinct wear-out failure modes. The first is the formation of an open circuit in a conductor, caused by the stripping away of metal ions by the high-intensity "electron wind."

The second is the etching of pits into the silicon, caused by its diffusion into the aluminum, and its subsequent transport away from the Si-Al interface. As the silicon is carried away, it is replaced by aluminum, and the process continues until a pit grows deep enough to short out an underlying junction.

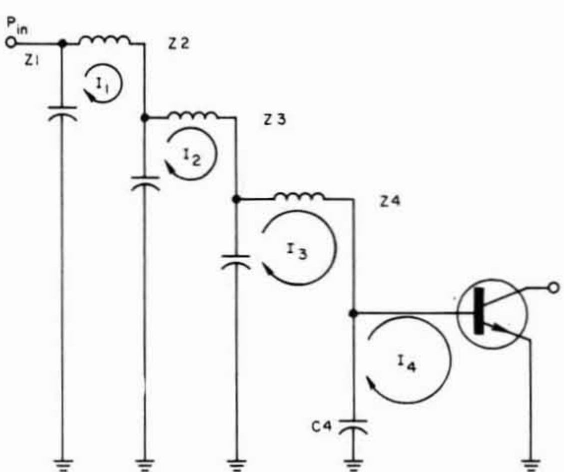
The transported silicon and aluminum are deposited where the conductor path widens—where the current density falls—and build up hillocks, whiskers and crystals that can then short out to other circuit points.

Metal ions can be more easily stripped away by the electron wind when they have been thermally activated enough to be essentially free of the metal lattice. The square of the current density determines the magnitude of the force acting on activated silicon ions. Thus the mass transport mechanism is dependent upon both current density and temperature.

For typical small conductor geometries (ignoring the increased rate of failure caused by a positive gradient in either current density, temperature or ion-diffusion coefficient), failures caused by mass transport become especially important when the temperature exceeds 150°C. By the time 200°C is reached, the lifetime can be measured in hours—provided, of course, that secondary breakdown doesn't strike first.



1. **Gain and efficiency are inversely related** below about 10 W, especially at low frequencies. These curves are based on the published data for over 100 devices made by 10 manufacturers. The curves are only a guide to typical behavior—they are not definitive for a specific transistor. Note that an efficiency reduction of about 25% can be expected when a high-gain, low-power device is compared with its high-powered counterpart.



2. **If C_4 is lossy, it can get hot enough to melt solder** because of the high circulating current (I_4) that flows in a low-impedance circuit. Since C_4 is intimately connected to the transistor, this temperature increase doesn't do the latter any good either.

For a driver collector efficiency of 40% and a need for 0.1 W of additional rf power, we'll have to put 0.1 W/0.40 = 0.25 more watts of dc through the device, leaving 0.25 − 0.10 = 0.15 W more power to be dissipated as heat.

The average junction-to-stud (or case) thermal resistance of the types of transistor under consideration is approximately 24.3°C/W. Thus, even if we assume that the transistor is operating on an infinite heat sink with zero thermal resistance between itself and the outside world, the temperature rise at the junction due to the increased dc dissipation is 0.15 W × 24.3°C/W = 3.64°C. This temperature rise in the driver could have been avoided by using lower-loss capacitors.

For a transistor with a maximum temperature rating of 200°C—and which we'll push to the accepted limit and operate at 180°C—we would discover that the hoped-for margin of 20°C had shrunk to 16°C. Add to this the fact that an infinite heat sink with zero thermal resistance is not ordinarily available, nor is a zero thermal resistance connection between the heat sink and the transistor, and we see even more of that safety margin disappear.

If getting out that last 0.5 dB forces the transistor into saturation, there will be a further decrease in stage efficiency, leading to an additional temperature rise. And all of this does not include the possibility that extra power will be demanded from the preceding stage because of the gain nonlinearity of the saturated transistor.

It should be noted here that the transistor-gain data of Fig. 1 is based on average—not minimum—values. In practice, of course, the engineer must face the prospect that he will receive a moderate number of minimum-gain devices along with the average ones. This is one further reason for exercising care in selecting external components that can contribute to losses. ■■

For bibliography, see Section 5-1 of this Handbook.

Test your retention

Here are questions based on the main points of this article. Their purpose is to help you make sure you have not overlooked any important ideas. You'll find the answers in the article.

1. What are two ways in which a lossy interstage capacitor can lead to transistor failure?

2. How are the efficiency, power rating and gain of rf transistors related?

3. Why is it desirable to mount the matching capacitor (C_4 in Fig. 2) so close to the transistor that mutual heating becomes a problem?

Evaluate transistor bandwidths

the easy way. Circuit Q's and interstage resistance levels provide a rough but rapid measure of available bandwidth.

Reprinted from Electronic Design, December 20, 1970

A lot of time can be wasted in trying to select a transistor for a wideband application. Typically, a tentative choice is made on the basis of gain, center frequency and power; then the engineer uses a Smith chart to see if the suggested transistor's immittances (Z_{in} and Y_{out}) when incorporated into a matching network, can provide the required bandwidth. If they can't, another tentative selection is made, and the process is repeated.

This can be quite tedious if several matching sections must be varied in an attempt to optimize the gain and VSWR over, say, an octave of bandwidth.

A better approach is to recognize that two major factors limit the available bandwidth of any transistor. These factors, which can be evaluated very quickly, are:

- The ratio of the transistor's series input reactance to input resistance, X_s/R_s.
- The quantity $[(R_1/R_2) - 1]^{1/2}$, where R_1 and R_2 are any two resistance levels that must be matched. For example, one can be a source resistance and the other the input resistance of the transistor. (When $R_1 \neq R_2$, R_1 is always the larger of the two.)

Series-to-parallel transformations

To see how the preceding factors affect transistor bandwidth, we must recall that any series resistance-reactance network can be replaced by a parallel equivalent (Fig. 1) at any specified frequency other than dc. At the frequency of their equivalence, the two circuits, naturally, have the same Q:

$$(X_s/R_s) = (R_p/X_p) = Q. \qquad (1)$$

The conversion equations are:

$$R_s = R_p/(Q^2+1) \qquad (2a)$$
$$X_s = R_s Q \qquad (2b)$$
$$R_p = R_s(Q^2+1) \qquad (2c)$$
$$X_p = R_p/Q. \qquad (2d)$$

The main facts that emerge from Eq. 2 are that R_p is always greater than R_s and that the reactances are always of the same type but will usually not be of the same value.

The quantity $X_s/R_s = Q$ imposes a fundamental limitation on a transistor because

$$(1/Q) = BW_{3\,dB}/f_o = \text{fractional bandwidth} \quad (3)$$

where $BW_{3\,dB}$ is the 3-dB bandwidth of the transistor's input circuit and f_o is the center frequency at which the measurements of X_s and R_s were made.

To see how the second quantity, $[(R_1/R_2 - 1]^{1/2}$, comes into play, imagine that the combination of the transistor's series input reactance, X_L, and its series input resistance, R_s, is lower than 50 Ω and that it must be matched to a 50-Ω source. If a suitable inductive reactance, X_L, is added in series with R_s and X_s it will build up the impedance to the desired level, $R_p = 50\ \Omega$, when the series combination is paralleled by a suitable value of capacitive reactance, X_c (Fig. 2). R_s and the series combination $X_L + X_s$ are transformed into the parallel combination R_p and X_p. The capacitive reactance is chosen so that $X_c + X_p = 0$ at f_o, thus making the circuit look like a pure resistance at the frequency of interest.

To see how this matching procedure affects the bandwidth, rewrite Eq. 2c to read

$$(R_p/R_s) = Q^2 + 1 \qquad (4)$$

or

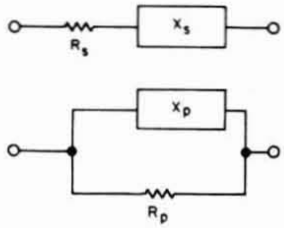
$$f_o/BW_{3\,dB} = Q = [(R_p/R_s) - 1]^{1/2}. \qquad (5)$$

This equation, which is very useful for practical, rapid, impedance matching makes clear that *the greater the difference in resistive levels to be matched, the narrower the bandwidth conveniently attainable.*

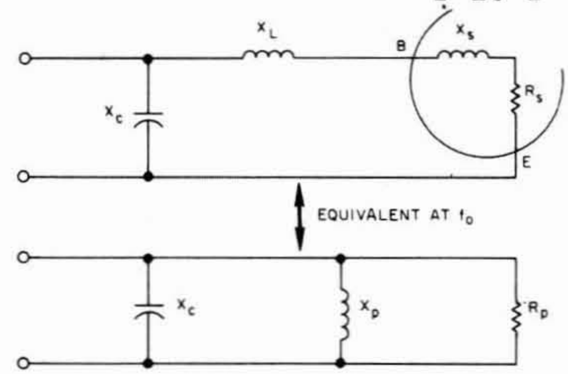
In deriving Eq. 5, it was assumed that R_s and X_s were so low that an additional reactance, X_L, had to be added to raise the impedance level. Actually, the inductance of the transistor leads often is too high, making it necessary to match up from the source to the transistor input, further restricting the bandwidth. Each matching section, of course, has its own fractional bandwidth and reduces the available bandwidth in proportion to it.

Since either sign of reactance may be used for matching, what is often done in this case is to add a capacitor (C_1) in series with the 50-Ω generator to raise its impedance up to the level of the transistor's input (Fig. 3a). Inductor L_1

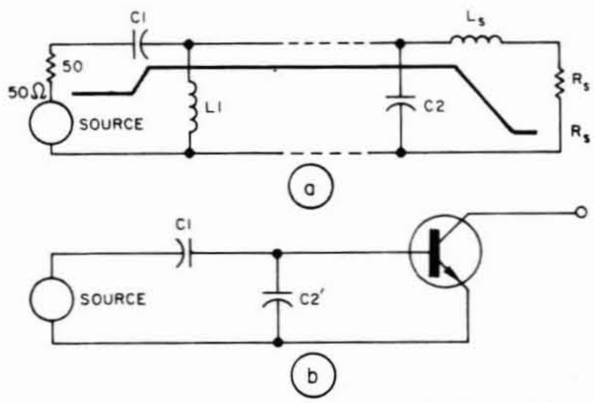
Vincent F. Perna, Vice President. Microwave Engineering, American Technical Ceramics, One Norden Lane, Huntington Station, N. Y. 11746.



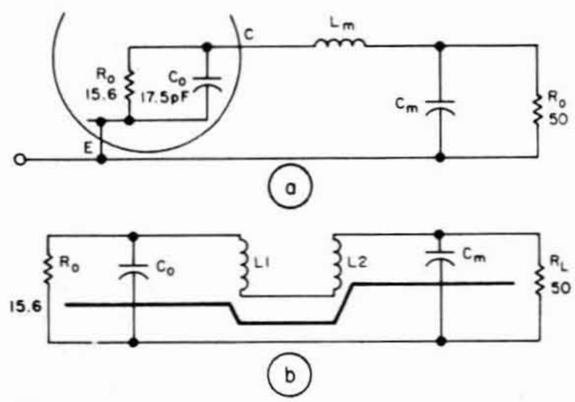
1. **At any specified frequency,** a series resistance-reactance pair can be represented by a parallel equivalent. The parallel resistance equivalent, R_p, will always be greater than the series resistance, R_s.



2. **Adding X_L raises the apparent input resistance** of the transistor (top) by increasing its effective parallel equivalent, R_p, (bottom). X_c is added to cancel the inductance at the operating frequency.



3. **The parasitic inductance compounds the problem** by making two impedance changes necessary (a). The colored line represents the impedance (to ground) of the various portions of the circuit. L_s raises the transistor resistance R_s above 50 Ω so that C_1 is needed to bring it down again. L_1 and C_2 can be combined, resulting in C_2' and a two-capacitor matching network (b).



4. **Output capacitance C_o can be integrated into a matching network (a)** at the output of a transistor. The easiest way to see how the circuit works is to imagine the inductor, L_m, to be broken into two series inductors, L_1 and L_2 (b). The colored line shows how the impedance is affected by the inductors. C_o cancels out the effect of L_1 at f_o; C_m does a similar job for L_2.

is used to resonate with C_1 at f_o and C_2 is used to similarly resonate L_s.

If the admittances of L_1 and C_2 are combined, a capacitor, C_2', usually results. This leads to the often-seen circuit of Fig. 3b.

Let's try a quick example

To see how easy it is to use these ideas, let's try to get a rough idea of the input circuit bandwidth available from a 2N3375. According to the manufacturer's data, taken at a collector current of 200 mA and a frequency of 300 MHz, the input impedance is 12.8 + j5.5 Ω. For a rough approximation we can neglect the 6-dB-per-octave gain slope and the variation of Q with frequency, and simply write: $Q = X_s/R_s = 5.5/12.8 = 0.43$. Thus the fractional bandwidth is $(1/Q) = 2.33$. At 300 MHz, this gives a bandwidth of 699 MHz, or a passband extending from dc to beyond 600 MHz.

All this assumes, of course, that a driver is available with a suitably low output impedance, conjugate over the full range. Alternatively, if we assume a purely resistive source, plugging into Eq. 2c, we see that a generator with a resistance of 14.7 Ω is needed.

Most likely, such a driver will not be available. Probably a 50-Ω unit will have to be used. As a result, the available bandwidth (using a single L-section matching network) would be reduced to 176 MHz, as can be verified by plugging into Eq. 5.

Don't forget the output circuit

So far, so good. But we have been dealing only with the transistor's input circuit. The output circuit will also restrict the bandwidth, and, unfortunately, output-resistance information is not always available from the manufacturer. It must either be measured, which is tedious, or calculated, which is quicker but less accurate since it involves several assumptions.

If high accuracy is not needed, the following equation can be used:

$$R_o = (V_{cc} - V_{sat})^2/2\ P_o. \qquad (6)$$

R_o is the output resistance, P_o is the output power, V_{cc} is the collector supply voltage and V_{sat} is the rf saturation voltage of the transistor (on the order of 3 V). Once Eq. 6 is solved, the result can be plugged into Eq. 5 for an estimate of the output-circuit bandwidth.

For example, for a 2N5637 at 20 W with $V_{cc} = 28$ V and $V_{sat} = 3$ V, Eq. 6 yields $R_o = 15.6\ \Omega$. Therefore, from Eq. 5, an initial estimate of the output-circuit bandwidth for a 50-Ω system operating at 300 MHz is 202 MHz.

Of course, the transistor's output really has a capacitance across the output resistance (Fig. 4a). This limits the bandwidth by affecting the circuit Q as described by Eq. 1. The larger the capacitance, the higher the circuit Q and the narrower the bandwidth.

Sometimes the shunt output capacitance is of such a value that it can be usefully employed in a simple matching network (Fig. 4a). A pi network is used to match the 15.6-Ω output resistance of the transistor to a 50-Ω line. The inductor L_m is theoretically equivalent to two series inductors, L_1 and L_2, which perform the match in two stages (Fig. 4b).

The two L-sections of Fig. 4b, of course, each contribute to bandwidth shrinkage, but the section represented by L_1 usually has a very low Q, and its effect on the bandwidth is therefore minimal.

As with other tightly coupled interactive networks, a wide variety of passband-gain combinations can be obtained through tuning trade-offs. For quick transistor evaluations, assuming synchronously single-tuned inputs and outputs, the expected over-all stage bandwidth (from Eq. 3) is approximately equal to:

$$BW_{3\ dB} = f_o/(Q_{input} + Q_{output}). \qquad (7)$$

Harmonic effects can be a problem

The usable portion of the available bandwidth is influenced by harmonic currents which can flow in the output circuit, lowering the output resistance and thus shrinking the bandwidth. (Some multi-octave circuits may, in fact, need to go push-pull to get rid of even-order harmonics.) Where efficiency is critical, circuit element values must be chosen to reflect an open circuit at the collector, even if the harmonics are out of the desired passband.

Thus, as broader bandwidths are required, space will have to be increased to accommodate the more elaborate matching networks that will be needed. Or, the engineer can reconsider his choice of transistor. ■■

BIBLIOGRAPHY:

See Section 5-1 of this Handbook

DON'T LET THE WRONG INPUT Q WRECK YOUR RF POWER AMPLIFIER

The very low impedances required in RF power amplifiers can cause high currents and excessive I^2R losses in passive matching components. Here is how to calculate what can be tolerated.

VINCENT F. PERNA, American Technical Ceramics

An RF power amplifier's input matching network must trade voltage for current as it lowers the signal impedance from the normal 50Ω line down to the 1 or 2Ω of the transistor input. This trade-off causes very large RF circulating currents to flow in the passive matching components. For example, capacitor C_3, in the widely used RF circuit of **Fig. 1**, can easily have a current of 1A or more flowing through it for 3W of input power. This, in turn, can cause I^2R losses in C_3 that can seriously degrade the amplifier gain and can even melt the solder that holds C_3 to the circuit. The problem is aggravated at frequencies over 100 MHz by the small physical size of the capacitor and the fact that the dissipation factors of most dielectrics increase with heat and frequency.

The designer, intent upon achieving optimum impedance matching between the 50Ω line and the transistor, often tends to overlook these large circulating currents and their heating effect. He may unintentionally have too high a loaded Q (Q_L) at the transistor input, which will increase the magnitude of this circulating current and the heating in the capacitor. Even if he does keep his Q within reason, he may unwittingly select a capacitor that has too high a series resistance.

What the designer needs is a quick method for looking at his circuit from this standpoint, and seeing quantitatively whether or not his design is within realistic limits—limits that are consistent with obtainable components, and which do not penalize the performance or the reliability of the stage.

Step-by-Step Procedure Simplifies Process

Here is a seven-step method for properly selecting the critical input capacitor, C_3, of the popular circuit configuration of **Fig. 1**.

(Continued)

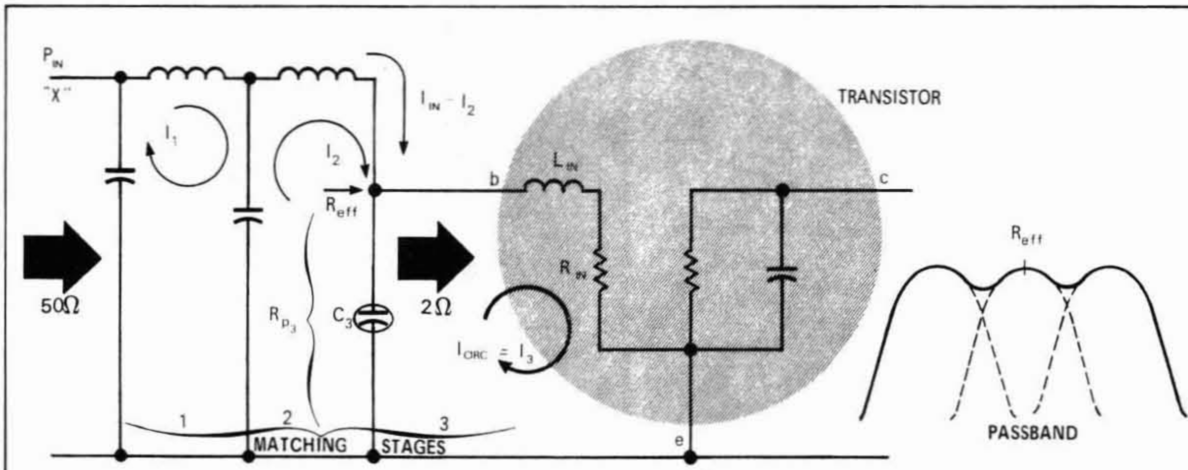


Fig. 1—**The high circulating currents** (I_{circ}) in the final matching stage of an RF amplifier can generate prohibitive I^2R losses in the passive components (particularly C_3 in this typical design). The designer must be able to calculate this current if he is to properly specify the critical components (C_3 in this example).

WRONG INPUT Q

(Continued)

Reprinted from EDN/EEE, November 15, 1971

Wrong Q (Cont'd)

The following circuit and device parameters are known at the start:

1. Circuit configuration . . . . as shown in **Fig. 1**.
2. Power in 3W
3. Allowable capacitor loss . . . . . . 0.05 dB (1.1%)
4. Operating center frequency, f_0 . . 300 MHz
5. Transistor input characteristics (at operating frequency) . . . . 2 + j5

$$R_{in} = 2\Omega$$

$$X_{L(in)} = 5\Omega$$

These values were picked to represent common practice.

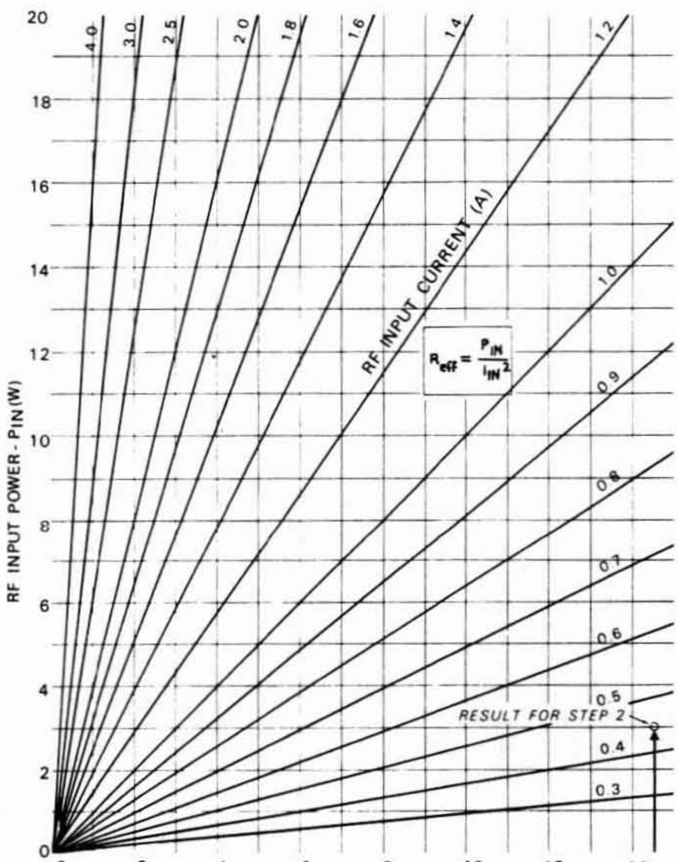


Fig. 2 - **Input current** (I_{in}) is plotted as a function of input power (P_{in}) and effective parallel input resistance (R_{eff}) on the final matching stage.

STEP 1 – Calculate R_{eff}, the effective parallel input resistance of the parallel-tuned network containing capacitor C_3 at the given center frequency.

The equation is:

$$R_{eff} = (Q_L^2 + 1)\, R_{in}$$

R_{in} is known because it is given for the transistor at the 300 MHz operating frequency and the circuit's bias conditions (not shown in **Fig. 1**). This R_{in}, of course, must be for the actual large-signal RF input with the device biased for handling the 3W signal. It is obtained from measurements, as any transistor would have to be operating in a highly nonlinear mode at this power level. Nowadays most RF transistor manufacturers supply this data on their device specification sheets.

The circuit load Q (Q_L) can be obtained from the relationship:

$$Q_L = \frac{X_L}{R_{in}} = \frac{5}{2} = 2.5$$

More will be said about the practical considerations for this parameter later.

Putting these values into the equation for R_{eff}:

$$R_{eff} = (2.5^2 + 1)\, 2 = (6.25 + 1)\, 2 = 14.5\Omega$$

STEP 2 – Find the input current, I_{in}, as defined in Fig. 1. This is known also as the line, or sustaining, current in RF circuits. Use the graph of **Fig. 2** for this. This graph solves the equation:

$$R_{eff} = \frac{P}{I_{in}^2}$$

where P, of course, is the signal power which has been given as 3W. Entering the graphs with these known values (as shown in **Fig. 2**), the RF input current, I_{in}, is found to be 0.46A.

STEP 3 – Find the circulating current in stage 3 by multiplying the RF input current found in the last step, STEP 2, by the loaded circuit Q (Q_L) found in STEP 1.

$$I_{circ} = I_{in} \times Q_L = 0.46 \times 2.5 = 1.15\text{A}$$

or, say, 1.2A. The values for these first three steps could also have been obtained from the graph of **Fig. 3**.

STEP 4 – Calculate the maximum permissible capacitor equivalent series resistance from the "allowed" or hoped-for I^2R dissipation:

$$R_s = \frac{P_{diss}}{I_{circ}^2}$$

The allowed I^2R dissipation, or P_{diss}, was given as 0.05 dB or 1.1%. This means the capacitor can dissipate only 33 mW. Therefore,

$$R_s = \frac{0.033}{(1.2)^2} = 0.023\Omega$$

STEP 5 – Calculate the capacitor's required minimum unloaded Q, Q_u.

$$Q_u = \frac{X_c}{R_s}$$

The capacitive reactance, X_c, required for C_3 at 300 MHz is given by:

$$X_c = \frac{R_{eff}}{Q_L} = \frac{14.5}{2.5} = 5.8\Omega$$

Therefore,

$$Q_u = \frac{5.8}{0.023} = 252$$

or, for a round figure, 300.

STEP 6 – Select a capacitor. Now that the allowable R_s is known, the designer can search for the component. The capacitance value would be found in the customary two steps:

$$X_c = \frac{R_{eff}}{Q_L} = \frac{14.5}{2.5} = 5.8\Omega = \frac{1}{2\pi fC}$$

$$C = \frac{1}{(6.28)\ (3\times10^8)\ (5.8)} = 91.5\ \text{pF}$$

Since present-day microwave power transistors often have dual emitter leads to minimize lead inductance, a sensible choice for C_3 would be to use two separate capacitors, one on each lead. This helps to achieve balanced currents on the two emitter leads and relaxes the specifications for the capacitors. Two 47-pF capacitors could then provide the 91.5 pF.

The designer must now have information on actual capacitors. He must either obtain this from the manufacturer or make his own tests on like devices. Fig. 4 represents the type of information available on one brand of microwave capacitors (device in this example is made by American Technical Ceramics). Entering the graph at the operating frequency and interpolating at 90 pF it is seen that the R_s is about .038Ω at 300 MHz.

Having the capacitors' measured R_s values plotted against capacitance helps the designer quickly evaluate the candidate and make trade-offs. For example, a single

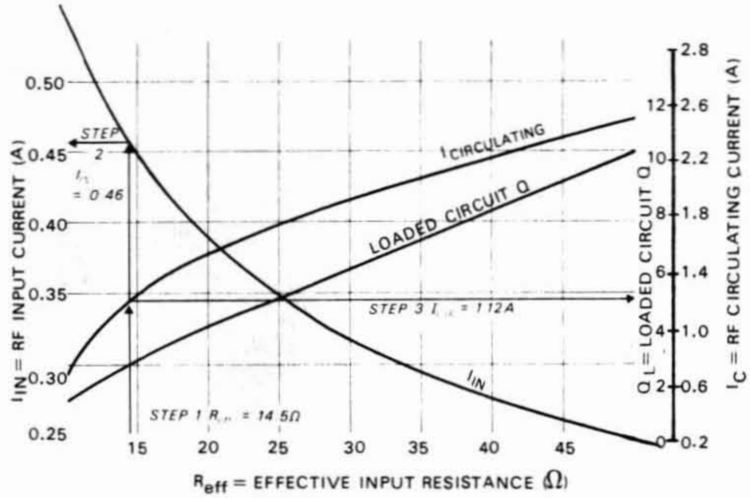


Fig. 3 – **Composite graph shows interrelationships between the parameters obtained in STEPS 1 through 3.** To obtain these plots, the transistor's lead inductance was held fixed and the base-emitter diode resistance was varied (see **Fig. 1** for model of transistor). Similar results can be obtained by holding the transistor resistance fixed and varying the base-emitter inductance of the base lead. These curves show very clearly how the "dangerous" circulating current goes up as Q_L goes up, substantiating the advice given in the text that Q_L should not be made too high.

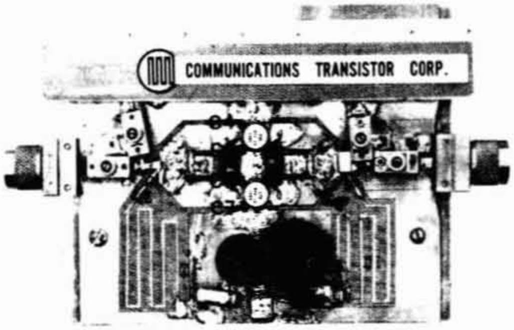
100 B 47 pF capacitor has an R_s of .055 . Two in parallel have a capacity of 94 pF and an R_s of only .027Ω. Since the computed allowable R_s from Step 4 was 0.023Ω, two 47 pF in parallel still fall a little short of the specified .023Ω. However, three 30 pF in parallel would meet the series resistance specification.

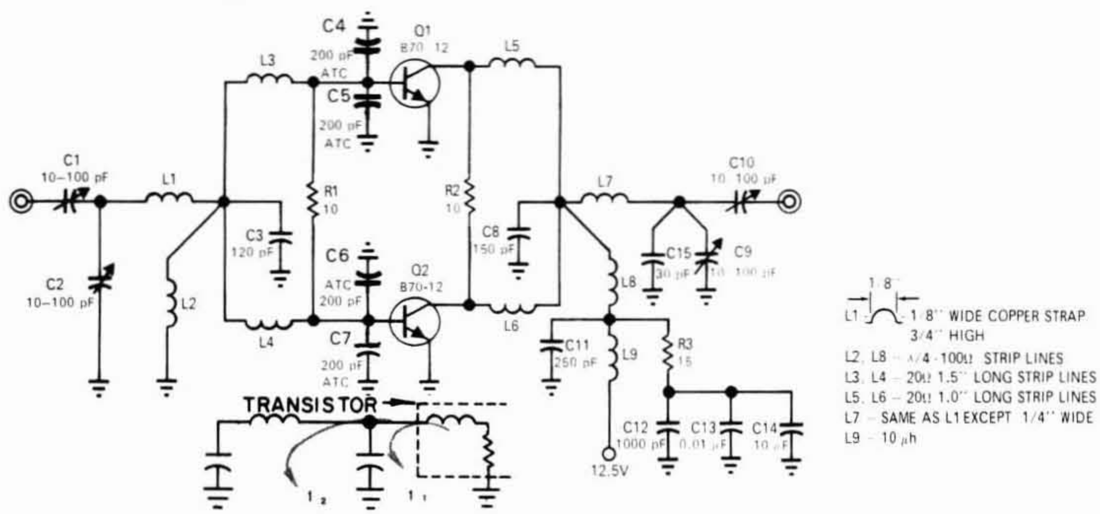
STEP 7 – Check temperature effects. The values for Q_u and R_s obtained from measurements – such as those plotted in **Fig. 4** – will only apply to a single temperature, which is most likely room temperature. But RF power amplifiers will rarely run at room temperature. They must be compact units able to operate within the restrictions of short wavelengths, and they must be shielded to prevent their RF from getting out and moisture from getting in; so it can be expected that they will run at some higher temperature – say, 125°C.

The temperature in the vicinity of the transistor will be even higher – perhaps 200°C. This heat will directly affect the tiny capacitors, which in typical RF designs will be placed snugly against the transistor.

Continued

A 140-W, 175-MHz Amplifier: Where Low-Loss Capacitors Are a Must





This land-mobile communications amplifier by Communications Transistor Corp., San Carlos, Calif., must use low-loss, low-parasitic inductance capacitors at the inputs to the two paralleled RF power transistors. The impedance going into the transistor is less than an ohm, so the circulating currents are several amperes, according to Joe Johnson, the CTC application engineer who designed the circuit.

For the first impedance matching step to be effective, the two shunt capacitors must present a capacitive reactance of 1Ω. If the capacitors used cannot present this low reactance, most of the current will flow in the shunt capacitors of other matching steps (i_2 instead of i_1). This causes the Q of the first effective matching step to double or triple, with a similar increase in circulating currents. Therefore, the loss will not only occur in the ineffective capacitor but in other matching components as well, because of the higher circulating currents. Losses caused by an ineffective first match may cause 3-dB loss in gain as well as saturated power output. Gain is too hard to come by at 140W and 175 MHz to throw it away in a passive component.

Johnson said that he used the low-loss "porcelain" types of chip capacitors made by American Technical Ceramics. These he has found to have better RF dissipations (higher Q_u's) and lower parasitic inductance than the more common barium titanate NPO types often used. (Vitramon and American Lava are said also to make low-loss porcelain units). Though these porcelain capacitors are about three times as expensive as the NPO types, their cost is still small compared to the expensive transistors needed to produce large powers at these frequencies. The chips cost about a dollar apiece in single quantities, while the transistors cost over $50 each in single quantities.

Incidentally, note the straightforward circuitry for paralleling the two power transistors. Johnson said that this eliminates the more exotic combiners, etc. often used to insure load-sharing at these frequencies. See Ref. 5 for more information.

The heat will work against the designer, for the Q_u will go down with increased temperature. This in turn will mean that the R_s will go up, and this regenerative action could lead to disasterous thermal runaway.

To appreciate the touchiness of the situation, consider that these tiny microwave capacitors have volumes on the order of one-thousandth of a cubic inch. Therefore, even though their dissipation might only be 30 mW or so, the power density could be as high as:

$$(30 \text{ mW})\,(1000) = 30 \text{ W/in}^3$$

This is close to the 40 W/in³ found at the tip of a soldering iron like an Ungar.

The only thing that may keep the capacitor's solder from melting may be the heat sinking provided by the typically flat-strip microwave conductors. The author recently heard of an engineer who thought he saw his capacitor chip moving, and indeed it was; the solder had melted and the device was being blown about in the solder puddle by the cooling fan! Obviously, an overheated input capacitor is not the nicest neighbor for an expensive RF transistor.

The objective of this final step, then, will be to estimate or make measurements that will forecast the deterioration of capacitor Q_u and R_s with temperature. ☐

Databank

1. "The Radio Engineering Handbook," By Keith Henny, McGraw-Hill, 3rd edition, 1941, p. 140.
2. "Communication Engineering," by William L. Everitt, Ph. D., McGraw-Hill, 2nd edition, 1937, p. 575.
3. "Electronic Engineers Handbook," by R. W. Landee, D. C. Davis, and A. P. Albrecht, McGraw-Hill, 1957, pp. 4-30.
4. "Evaluate Transistor Bandwidths the Easy Way," by V. F. Perna, Electronic Design Magazine 26, Dec. 20, 1970, pp. 40-42.
5. "175 MHz, 12V, 140W Amplifier," Application Note 2.1.8.4C, Communication Transistor Corp., San Carlos, Calif.
6. "Radiotron Designer's Handbook," edited by F. Langford-Smith, published by Radio Corp. of America, 4th edition, 1953, p. 410.

Vince Perna is a RF engineer who got so involved in helping one of his component suppliers that he ended up as vice president for applications with that company—American Technical Ceramics. Vince is impressed by the thirst for RF know-how that exists among today's engineers. He finds that his basic half-hour application lectures at RF houses often wind up with 2-hr question-and-answer sessions.

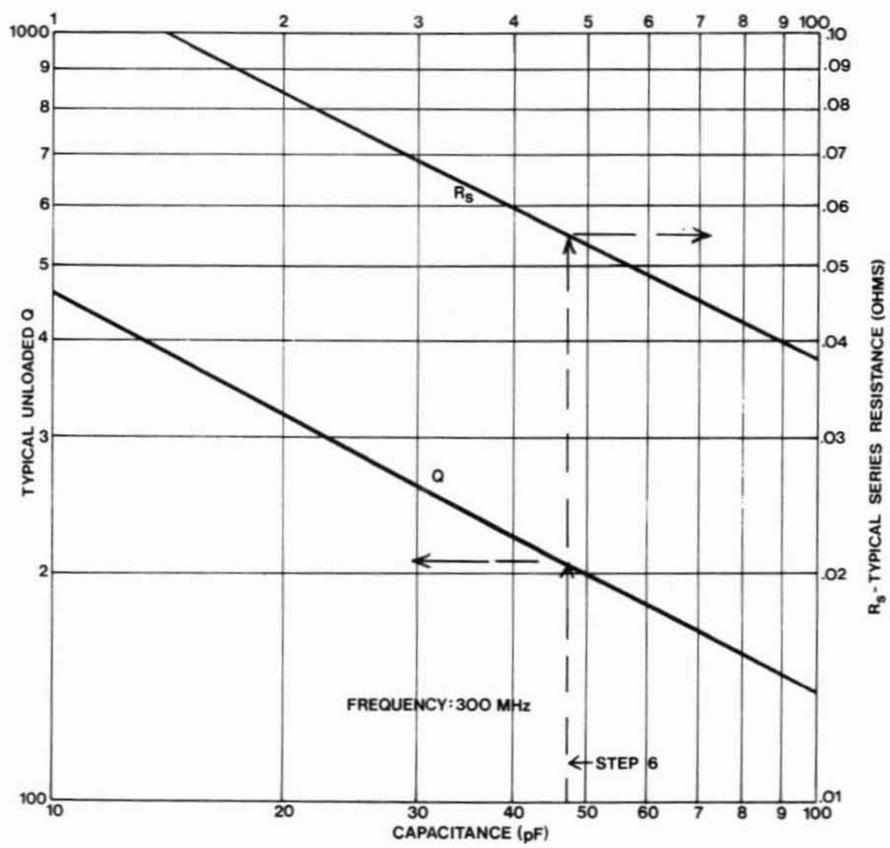


Fig. 4—**Appropriate data on available chip capacitors** is required if the designer is to select an RF unit that will have sufficiently low series resistance at the operating frequency. The data shown is for American Technical Ceramics ATC100-B chip capacitors.

HOW TO CALCULATE REQUIRED CAPACITOR Q FROM DESIRED MATCHING NETWORK EFFICIENCY OR LOSS

by Vincent F. Perna
Vice President
Microwave Engineering
American Technical Ceramics

As discussed in Handbook Section 2, the power transfer efficiency for a double-loaded, single L-C section with VSWR conditions of $S_{11} \neq 0$ and $S_{22} \neq 0$ is as given below in equation (1):

$$E_{ff}\ (\text{power}) = (\eta) = [1 - \left(\frac{Q_L}{Q_u}\right)]^2 \qquad (1)$$

More convenient to use is its inverse expression in equation (2) in terms of loss (in dB):

$$\text{Loss (dB)} = 10 \log_{10} [1/\eta] = 20 \log_{10} [Q_u/(Q_u - Q_L)] \qquad (2)$$

This article derives these equations, and, through charts plotted from them, presents a rapid means of determining required capacitor Q.

From equation (2) we may determine:

(A) *the required capacitor unloaded Q,* when the circuit bandwidth, inductor unloaded Q, and allowed network loss are known;

(B) *the required inductor unloaded Q,* when the maximum available capacitor unloaded Q, the desired maximum circuit loss, and the bandwidth are known;

(C) *the circuit loss we are forced to accept* with commercially available components when a given bandwidth is required.

A typical circuit application is shown in Figure 1:

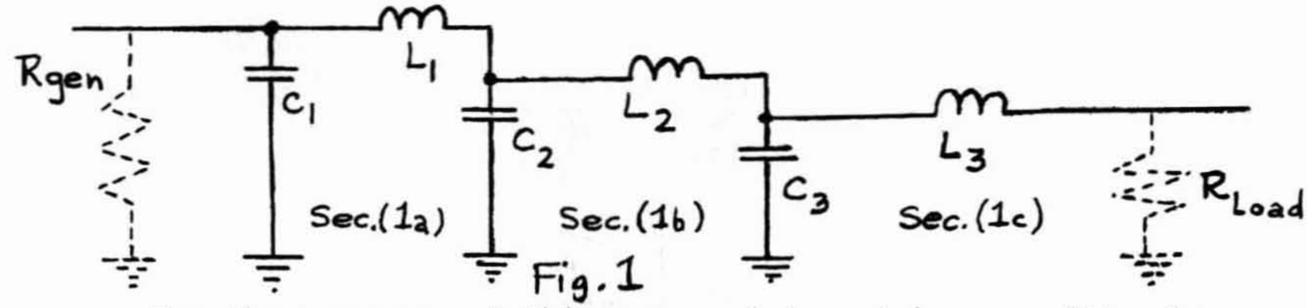


Fig. 1

For the purposes of this paper, let us take case (A), above. The power transfer efficiency (in percent) as a function of circuit unloaded-to-loaded Q ratio is shown in Figure 2. At first glance at the slope of Figure 2, one might assume that the 90 to 93.% efficiency level for each LC section was roughly the point of diminishing returns on Q_u-to-Q_L ratio.

However, when we compare this (Q_u/Q_L) of 20 with its expression in power loss in decibels (Figure 3):

$$\text{Loss (dB)} = 10 \log_{10}\left[\frac{1}{\eta}\right] = 20 \log_{10}\left[\frac{Q_u}{Q_u - Q_L}\right]$$

we find this is actually an intolerably poor ratio. (The loss, in this case, being 0.46 dB for just one section).

This becomes especially clear when we consider that RF power transistors may have only 3 to 8 dB gain, and may require several such network sections between them and the outside world to attain adequate bandwidth (e.g. driver transistor, a transmission line, or the ultimate load directly).

If we have decided upon an overall acceptable loss for a complete network, and from experience have a feeling for the relative loss distribution throughout the network, we can then specify the required maximum loss per section.

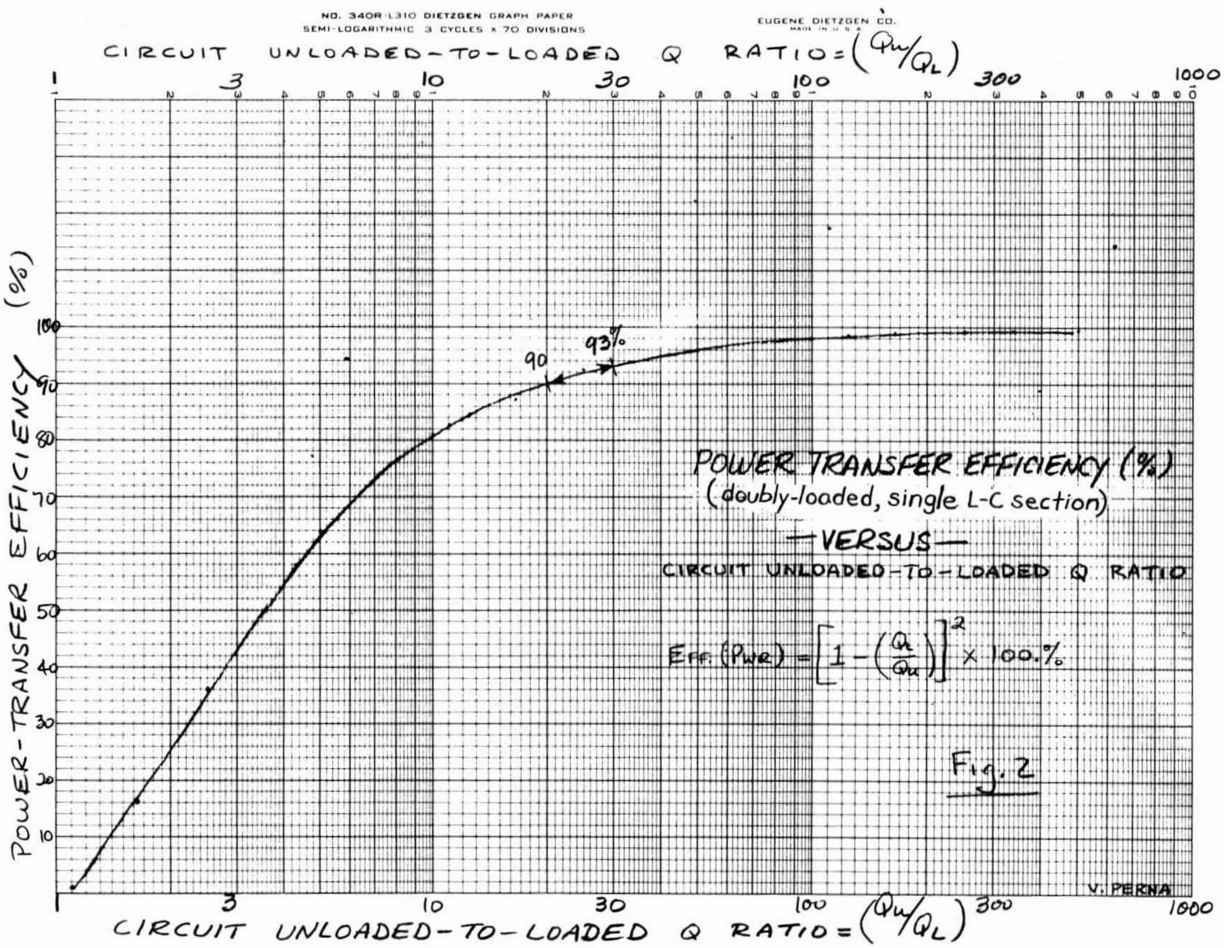

NO. 340R-L310 DIETZGEN GRAPH PAPER
SEMI-LOGARITHMIC 3 CYCLES X 70 DIVISIONS
EUGENE DIETZGEN CO.
MADE IN U.S.A.
CIRCUIT UNLOADED-TO-LOADED Q RATIO = (Q_u/Q_L)
1
3
10
30
100
300
1000
POWER-TRANSFER EFFICIENCY (%)
100
90
80
70
60
50
40
30
20
10
90
93%
POWER TRANSFER EFFICIENCY (%)
(doubly-loaded, single L-C section)
—VERSUS—
CIRCUIT UNLOADED-TO-LOADED Q RATIO
$\text{Eff.}(P_{WR}) = \left[1 - \left(\frac{Q_L}{Q_u}\right)\right]^2 \times 100\%$
V. PERNA
CIRCUIT UNLOADED-TO-LOADED Q RATIO = (Q_u/Q_L)

Fig. 2

Handbook Section 2-18 indicates that in some 3-section impedance transformers of the type shown in Figure 1, we may experience increasing losses through network sections (A), (B), and (C) approximately in the ratio of 1:2:4, respectively, due to the rapid fall of component Q's with circuit impedance level (and rising RF current level.)

Thus, if at some frequency we desire that the total network loss is not to exceed 0.5 dB, then the losses in section (1a) would have to be $\left(\frac{1}{1+2+4}\right) = \left(\frac{1}{7}\right)$ or 14.3% of that. Similarly, the losses in section (1b) would be $\left(\frac{2}{7}\right)$ or 28.6% of 0.5 dB, and in section (1c) 57.2%. Thus for sections (1a), (1b), and (1c) we would require losses no greater than 0.07, 0.15, and 0.29 dB, respectively.

Let us assume, that losses in C_2L_2 should be 0.15 dB max, and see what capacitor Q this will require.

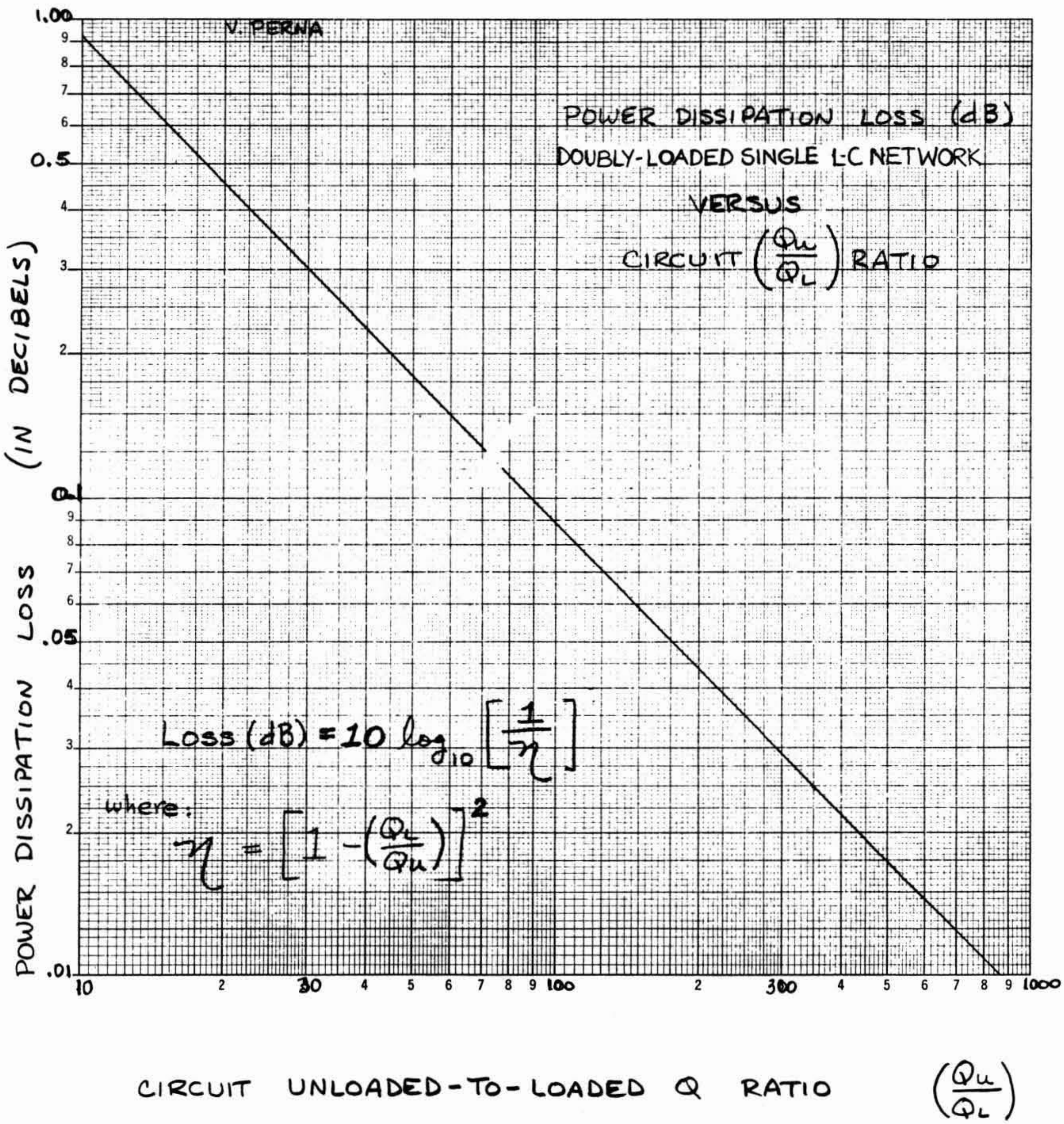
From Figure 3, we see this necessitates that $(Q_u/Q_L) = 60$. To determine Q_u, we must know the bandwidth of the LC section. This is dictated by the design method, and in this case let's assume the BW is 66.% ($Q_L = 1.5$). (See Reference 7.)

The unloaded Q of section (1b) must then be:

$$Q_u = 60\ Q_L = 60.\ X\ 1.5 = 90.$$

If we were to assume there were no losses in the capacitor, this would mean an inductor Q of 90 was needed. At UHF, however, capacitors can have such serious losses that circuit performance is considerably degraded, even though the inductor is usually the major source of loss.

The question then arises as to what must the capacitor-to-inductor loss relationship be?



Logarithmic, 2 x 2 Cycles

Figure 3

Intuitively, we sense that if the capacitor's loss were only a few percent of that of the associated inductor, the loss due to the inductor would be a sufficient approximation to the required LC-section loss.

Figure 4, in fact, displays this interrelationship exactly, and was developed (see Appendix II) from:

$$(1/Q_u) = (1/Q_{cap}) + (1/Q_{ind})$$

which becomes (see Appendix IV):

$$\left(\frac{Qu}{Qind}\right) = \left(\frac{y}{1+y}\right) \quad \text{where} \quad y = \left(\frac{Qcap}{Qind}\right)$$

Suppose now that the inductor Q consistently attainable in the proposed circuit layout turns out to be only slightly higher (say, 3.%) than the unloaded Q required of tuned section (1b) as a whole. This forces us to specify a capacitor Q value that will result in an unloaded circuit Q that is degraded by the capacitor to no more than 3.% below the level available from the inductor alone.

From Figure 4 we can see, however, that to attain only 3.% degradation would require a capacitor with a Q of 32 times that of the inductor. This may be a problem, depending upon frequency and physical configuration.

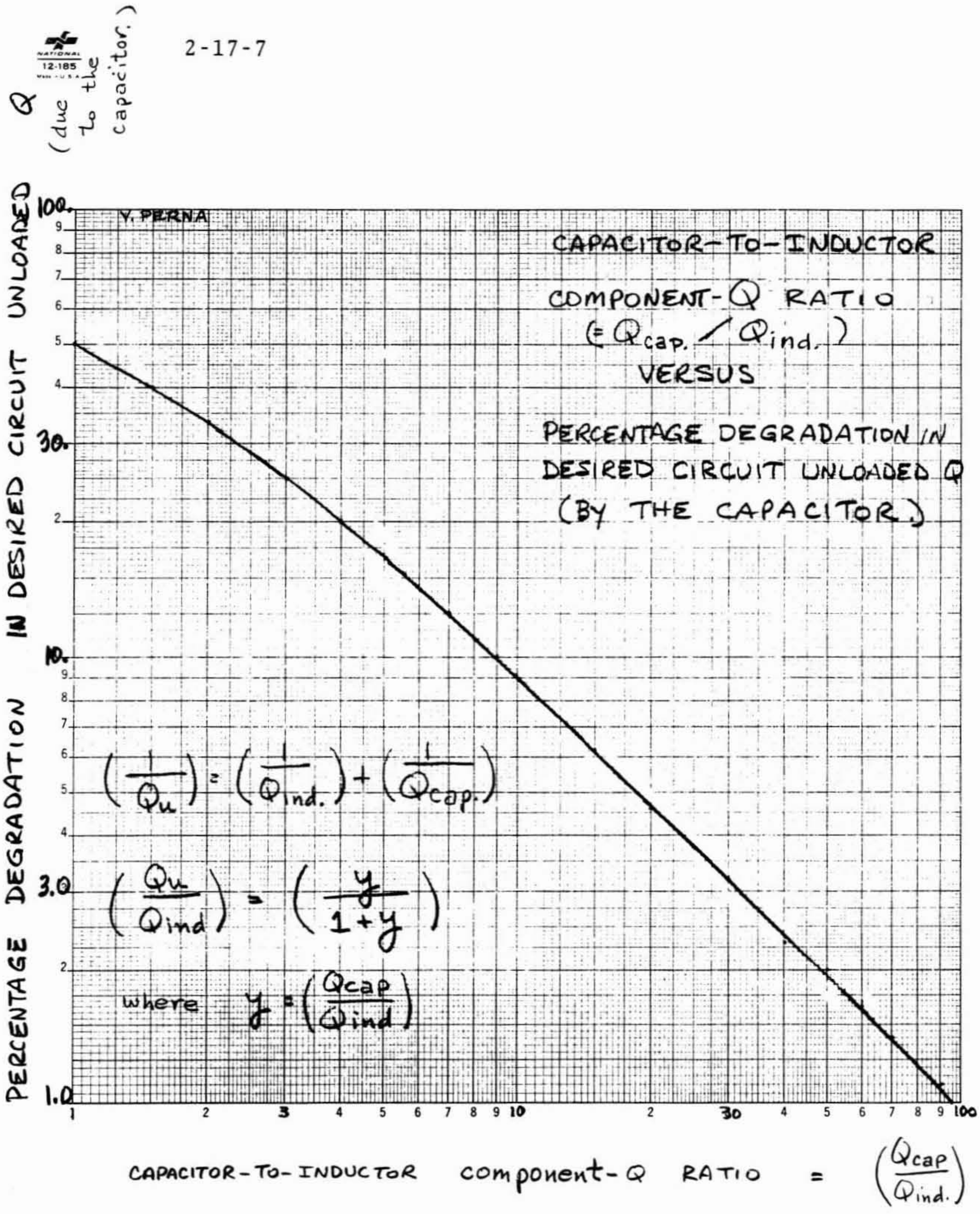
Once the required unloaded circuit Q and the (Q_{cap}/Q_{ind}) ratio that gives the allowed degradation are known, the necessary capacitor unloaded Q may be determined from Figure 5. This graph was derived from a reorganization of the preceding equation to be:

$$Q_{capacitor} = Q_u\left[1 + \left(\frac{Q_{capacitor}}{Q_{inductor}}\right)\right]$$

wherein Q_{cap} was solved for various ratios of (Q_{cap}/Q_{ind}) for specific values of circuit Q_u.

Thus, in our circuit example with a Q_u of 90 and a (Q_{cap}/Q_{ind}) ratio of 32, we see from Figure 5 that Q_{cap} must be about 2900.

2-17-7



Logarithmic, 2 x 2 Cycles

Figure 4

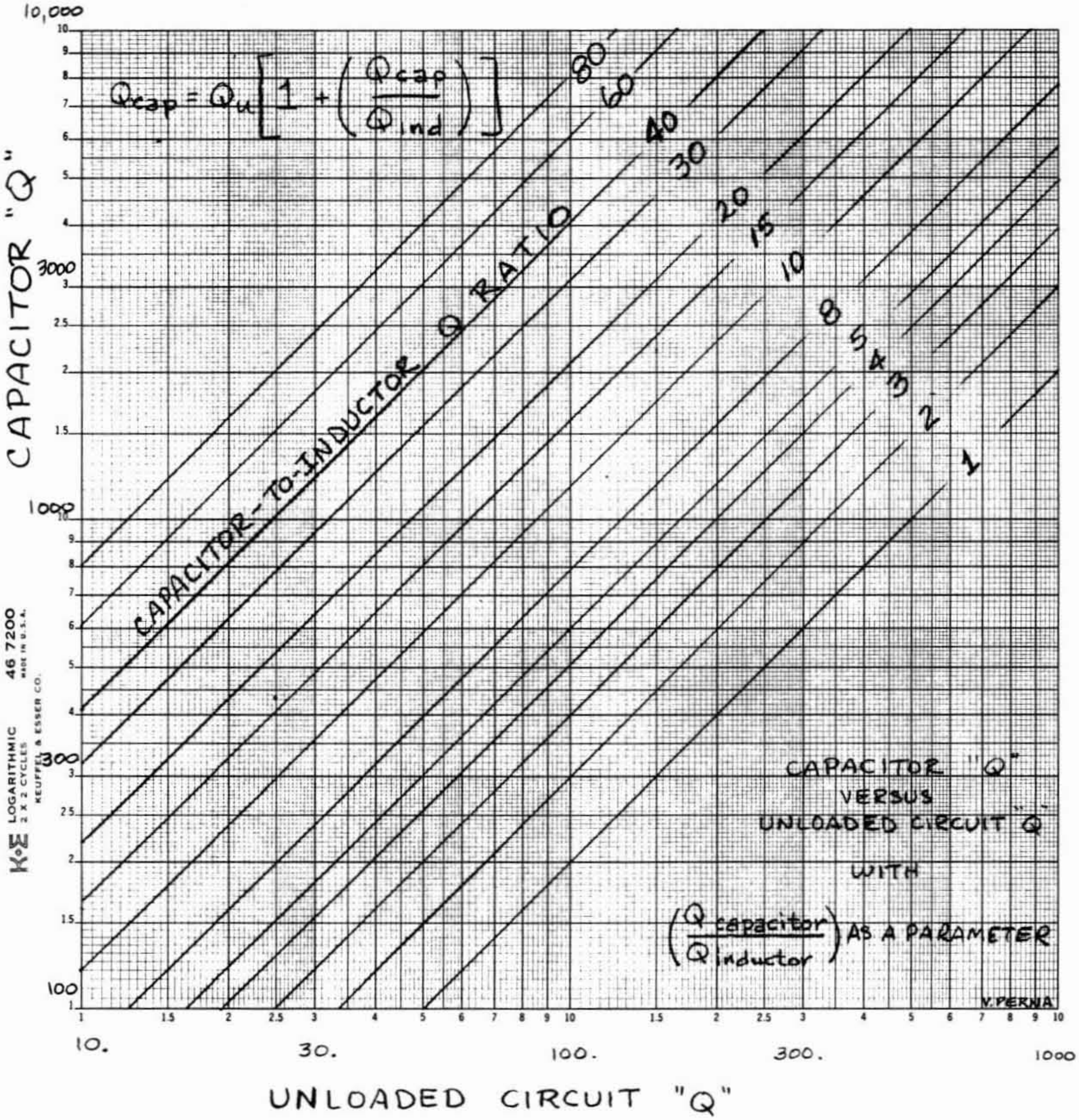


Figure 5

CAPACITOR Q FROM NETWORK EFFICIENCY

Practical Limitations on the Method:

For multi-layer capacitors in the UHF range, very high Q turns out to be acceptable only for small values of capacitance. For example, if we happen to be talking about values of 10 to 50. pF in the several hundred Megahertz frequency range, we will face Q's of about 3000 to 250, respectively (see Reference 12.)

As a consequence, because of these limitations in commercially available components, simply desiring a certain level of dissipation loss in a network does not mean that we shall experience it in practice.

Where RF power levels are low (hence the impedance levels relatively high), we are more likely to attain on the bench what we desire based on theoretical considerations.

If, on the other hand, we try putting significant RF current through a network, we may be rudely surprised to find that our losses are much higher than anticipated. The problem is that I^2R heating causes component Q to decrease, resulting in increased losses. This Q degradation-versus-temperature is a function of RF current level, ambient temperature, the Watts-dissipated-per-unit-volume, mounting method, heat removal rate, and thermal resistance. (See Appendix III.)

Furthermore, additional power loss caused by detuning may be experienced due to significant capacitance shift at temperatures beyond some manufacturers' rated limits. These temperatures can easily occur when capacitors are carrying high RF current and are mounted close to a heat source inside equipment which is subjected to an outside ambient temperature of 125°C.

Thus, although it is valuable to ask: "What component Q will be necessary, based upon some maximum desired circuit loss?", much time-wasting backtracking and frustration may be avoided by supplementing these preliminary calculations with an examination of what the circuit losses will be based upon available component Q's.

To have confidence in the results requires a comparative examination among available capacitor types for their Q-versus-temperature and temperature-versus-RF current. Once this is known, the circuit loss may be determined in advance by employing Section 2-18 of this Handbook.

BIBLIOGRAPHY

(1) Transistor Circuit Analysis and Design, by F.C. Fitchen, Van Nostrand, 1960, pages 249-250.

(2) Radio Engineer's Handbook, by F.M. Terman, McGraw-Hill, 1943, page 450.

(3) Microwave Transmission Design Data, by T. Moreno, Sperry Gyroscope Co., Inc. Publication #23-80, May, 1944, page 205.

(4) Very High Frequency Techniques, Harvard Radio Research Lab., Boston Technical Publishers, 1965, page 614 and 745.

(5) Reference Data for Radio Engineers, Fourth Edition, ITT Corp. page 579 and 122.

(6) "How to Determine Required Capacitor Q From Allowed Capacitor Power Loss," (RF Capacitor Handbook Section 2-16); also: "Don't Let the Wrong Input Q Wreak Your RF Power Amplifier," by V.F. Perna, EDN/EEE, Nov. 15, 1971 pp. 31-35.

(7) "Evaluate Transistor Bandwidth the easy way," by V. Perna, Electronic Design, Dec. 20 1970 pp. 40-42; (RF Capacitor Handbook Section 2-15.)

(8) "Have You Wondered Why Your Competition Matches Your Performance With Fewer Stages?", (RF Capacitor Handbook Section 2-7).

(9) Circuit Design for Audio, AM/FM, and TV, McGraw-Hill Book Co., Texas Instrument Electronic Series, 1967, pages 113, 132 & 133.

(10) Modern Transistor Electronics Analysis and Design, by F.K. Manasse, J.A. Ekiss, and C.R. Gray, Prentice Hall, 1967, pages 144-150.

(11) Communication Engineering, by W.L. Everitt, McGraw-Hill, 2nd Edition, 1937, pages 260-262.

(12) "Chip Capacitor Dielectric Effects on Hybrid Microwave Amplifiers," by V.F. Perna, Proc. of 1971 Int'l. Microelectronic Symposium (ISHM, Oct. 13th, Chicago, Ill.) pages 8-5-1 thru 8-5-18; (RF Capacitor Handbook Section 2-5).

(13) "On Limitations of Broadband Impedance Matching Without Transformers," by S. Plotkin and N.E. Nahi, IRE Trans on Circuit Theory, June, 1962, pages 125-132.

(14) "Theoretical Limitations on the Broadband Matching of Arbitrary Impedance," by R.M. Fano, Journal of the Franklin Institute, January, 1950, pages 57-83. Also: Technical Report No. 41, January 2, 1948, Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.

(15) "A Tutorial Exposition of some Key Network-Theoretic Ideas Underlying Classical Insertion-Loss Filter Design," by D.C. Youla, Proc. of IEEE, Vol. 59, No. 5, May, 1971, pages 760-799.

(16) "Synthesis of Reduced-Ripple Tschebyscheff Filters," by R.D. Wanselow, The Microwave Journal, July, 1971, pages 36-42.

(17) "Insertion Loss Concepts," by R.W. Beatty, Proc. IEEE, June, 1964, pages 663-671.

(18) "Evaluating the Impedance Broadbanding Potential of Antennas," by A. Vassiliadis and R.L. Tanner, IRE Transactions on Antennas and Propagation, July, 1958, pages 226-231.

(19) "Output Networks for Radio-Frequency Power Amplifiers," by W.L. Everitt, Proc. of IRE, Vol. 19, No. 5, May, 1931, pages 725-737.

(20) "A Note on the Design of Series and Parallel Resonant Circuits," by L.E. Q. Walker, Marconi Review, No. 63, December, 1936, pp. 7-11.

(21) "Efficiency of Inductive Coupling," by A.C. Hudson, Electronics Magazine, December, 1947, page 138.

(22) Antenna Engineering Handbook, H. Jasik, Editor, McGraw-Hill Book Co. First Edition, 1961, pages: 1-4, 19-3, 2-46 to 2-50, 25-4, 28-7, 31-3, 34-22, 35-16 to 35-18.

(23) "Broadband Reflectionless Matching with Minimum Insertion Loss," by H.J. Carlin and R. LaRosa, Proc. of Symposium of Modern Network Synthesis, New York, N.Y., April 16-18, 1952, pages 161-178.

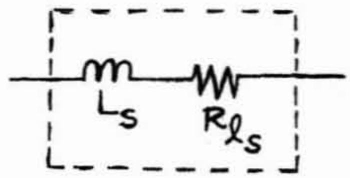
(24) "Single- and Coupled-Circuit Systems," by E.S. Purington, Proc. of IRE, Vol. 18, No. 6, June, 1930, pages 983-1016.

(25) "Generalized Characteristics of Linear Networks," by E.K. Sandeman, The Wireless Engineer, December, 1936, pages 637-646.

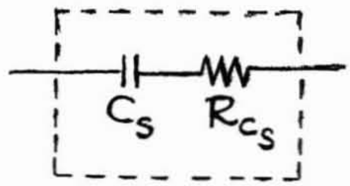
APPENDIX I

<u>POWER TRANSMISSION LOSS (in dB) OF ONE, DOUBLY-LOADED (generator + load) IMPEDANCE-MATCHING LC SECTION AS A FUNCTION OF LOADED-TO-UNLOADED CIRCUIT Q RATIO.</u>

<u>Component Q:</u>

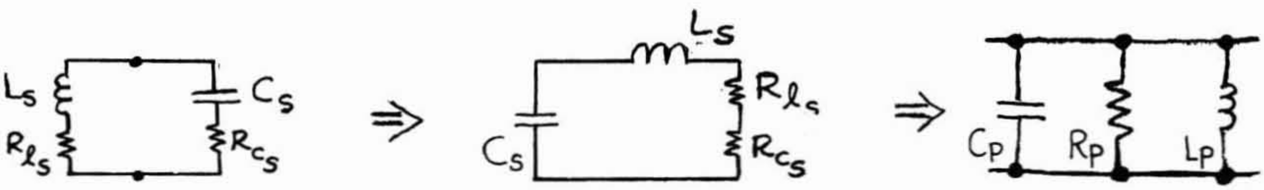
Inductor: 

$Q_{ind} = (X_{L_S}/R_{\ell_S})$

Capacitor: 

$Q_{cap} = (X_{C_S}/R_{C_S})$

<u>Unloaded Circuit Q at Resonance:</u>



---where R_p is the parallel equivalent of the series combination of R_{ℓ_s} and R_{C_s} and from using the relationships:

$$\left(\frac{X_{L_S}}{R_{\ell_S} + R_{C_S}}\right) = \left(\frac{X_{C_S}}{R_{\ell_S} + R_{C_S}}\right) = Q_u = \left(\frac{R_P}{X_{L_P}}\right) = \left(\frac{R_P}{X_{C_P}}\right)$$

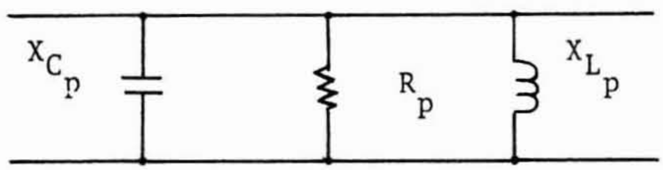
$$R_P = R_{S(\text{total})} \; (1 + Q_u^2) \qquad \text{where } R_S = R_{\ell_S} + R_{C_S}$$

and from Bibliography Reference (5) on page 2-17-10 of this Handbook, where:

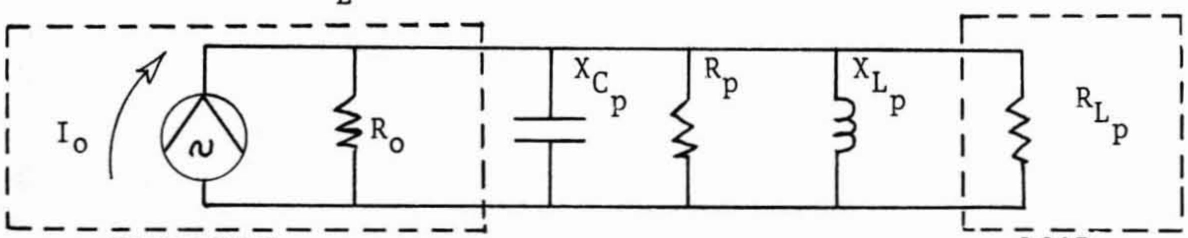
$$X_{L_p} = X_{L_s}\left[1+\left(\frac{1}{Q_u}\right)^2\right] \text{ and } X_{C_p} = X_{C_s}\left[1+\left(\frac{1}{Q_u}\right)^2\right]$$

Doubly-Loaded Circuit Q (= Q_L):

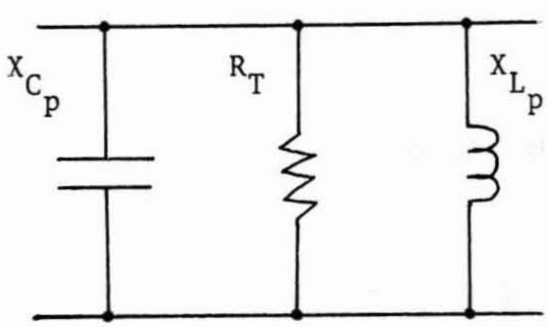
If the basic tuned-circuit (with all self-resistances and reactances converted to a parallel representation):



is then loaded on both "ends" by being driven by a (purely resistive) current generator and connected to a (purely resistive) load, R_L:

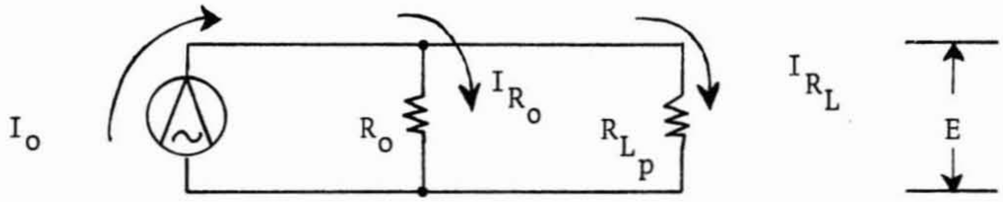


the effect upon Q of these two additional shunt current paths must be added to that of the original R_p due to L and C alone. Combining R_o, R_p, and R_{L_p} into a total shunting resistance designated by R_T, we can then determine the doubly-loaded circuit Q at resonance:



$$Q_L = \left(\frac{R_T}{X_{C_p}}\right) = \left(\frac{R_T}{X_{L_p}}\right)$$

Power Transfer at Resonance: theoretical, ideal case where $R_p = \infty$, so there is no loss in the parallel-tuned circuit reactances.



$$\left(\frac{1}{R_o} + \frac{1}{R_{L_p}}\right) = \frac{1}{R^*} = G^* = \left(G_o + G_{L_p}\right)$$

Sharing of current is given by:

$$I_{R_{L_p}} = I_o\left(\frac{G_{L_p}}{G_o + G_{L_p}}\right) = I_o\left(G_{L_p}\right)\left(\frac{1}{G_o + G_{L_p}}\right)$$

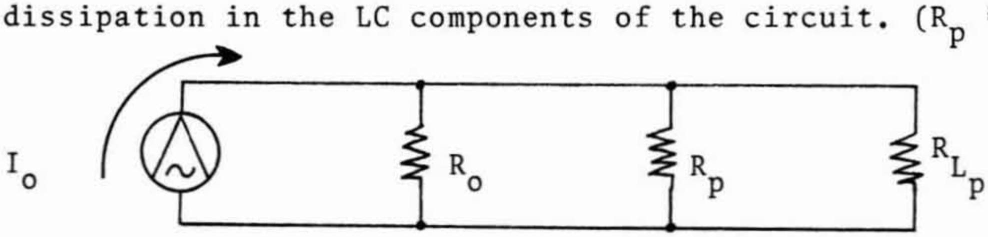
which, in terms of circuit resistances, becomes

$$I_{R_{L_p}} = I_o\left(\frac{1}{R_{L_p}}\right)\left(R^*\right) = \frac{I_o R^*}{R_{L_p}}$$

Thus, the power absorbed in the load in a doubly-loaded circuit with an ideal, theoretical, lossless LC network is:

$$P_{R_{L_p}} = \left(I_{R_{L_p}}\right)^2 R_{L_p} = \left(\frac{I_o R^*}{R_{L_p}}\right)^2 R_{L_p} = \frac{(I_o R^*)^2}{R_{L_p}}$$

Power Absorbed in the Load (R_{L_p}) in the Doubly-Loaded, Real-Life, Non-Ideal Case where there is some power lost due to dissipation in the LC components of the circuit. ($R_p \neq \infty$).



$$\left(\frac{1}{R_o} + \frac{1}{R_p} + \frac{1}{R_{L_p}}\right) = \frac{1}{R_T} = G_T = \left(G_o + G_p + G_{L_p}\right)$$

Sharing of current is given by

$$I_{R_{L_p}} = \left(\frac{I_o \; G_{L_p}}{G_o + G_p + G_{L_p}}\right) = I_o \left(G_{L_p}\right)\left(\frac{1}{G_o + G_p + G_{L_p}}\right)$$

which, in terms of resistances, is:

$$I_{R_{L_p}} = I_o \left(\frac{1}{R_{L_p}}\right)\left(R_T\right) = \left(\frac{I_o \; R_T}{R_{L_p}}\right)$$

Thus, the power absorbed in the load in a doubly-loaded, real life, lossy LC circuit is:

$$P_{R_{L_p}} = \left(I_{R_{L_p}}\right)^2 R_{L_p} = \left(\frac{I_o \; R_T}{R_{L_p}}\right)^2 R_{L_p} = \frac{\left(I_o \; R_T\right)^2}{R_{L_p}}$$

The Power Transfer Efficiency : Ratio of Real-Life to Ideal-Case Powers Absorbed in the Load: symbolized by

$$\frac{P_{R_{L_p}} \text{ (real-life)}}{P_{R_{L_p}} \text{ (ideal case)}} = \frac{\left[\dfrac{(I_o R_T)^2}{R_{L_p}}\right]}{\left[\dfrac{(I_o R^*)^2}{R_{L_p}}\right]} = \frac{\left[R_T\right]^2}{\left[R^*\right]^2}$$

$$= \frac{\left[\dfrac{1}{G_o + G_p + G_{L_p}}\right]^2}{\left[\dfrac{1}{G_o + G_{L_p}}\right]^2} = \left(\frac{G_o + G_{L_p}}{G_o + G_p + G_{L_p}}\right)^2$$

However, the unloaded circuit Q ($=Q_u$) in the real-life case is given by $Q_u = (R_p / X_{L_p}) = (R_p / X_{C_p})$ at resonance, and thus

$$X_{L_p} = (R_p / Q_u)$$

Similarly the loaded circuit Q ($=Q_L$) in the real-life case is given by:

$$Q_L = (R_T/X_{L_p}) = (R_T/X_{C_p})$$

Consequently

$$Q_L = \left[\frac{1}{(G_o + G_p + G_{L_p})\, X_{L_p}}\right]$$

and therefore

$$Q_L = \left[\frac{1}{(G_o + G_p + G_{L_p})\left(\frac{R_p}{Q_u}\right)}\right]$$

solving this above equation for $(G_o + G_{L_p})$:

$$Q_L\ (G_o + G_p + G_{L_p}) = (Q_p/R_p)$$

$$(G_o + G_p + G_{L_p}) = Q_u/(Q_L R_p)$$

therefore: $(G_o + G_L) = \left(\dfrac{Q_u}{Q_L\ R_p}\right) - G_p = \left(\dfrac{Q_u}{Q_L R_p}\right) - \left(\dfrac{1}{R_p}\right)$

and:

$$(G_o + G_{L_p}) = \left(\frac{1}{R_p}\right)\left[\left(\frac{Q_u}{Q_L}\right) - 1\right] = \left(\frac{1}{R_p}\right)\left[\left(\frac{Q_u}{Q_L}\right) - \left(\frac{Q_L}{Q_L}\right)\right]$$

$$= \left(\frac{1}{R_p}\right)\left(\frac{Q_u - Q_L}{Q_L}\right) = \left(\frac{Q_u - Q_L}{R_p\ Q_L}\right)$$

Then, substituting this plus:

$$G_o + G_p + G_{L_p} = Q_u/(Q_L R_p)$$

in the expression for the power-transfer efficiency:

$$\frac{P_{R_L\ (\text{real life})}}{P_{R_L\ (\text{ideal case})}} = \left(\frac{G_o + G_{L_p}}{G_o + G_p + G_{L_p}}\right)^2 = \frac{\left(\dfrac{Q_u - Q_L}{R_p Q_L}\right)^2}{\left(\dfrac{Q_u}{Q_L R_p}\right)^2}$$

$$\left[\left(\frac{Q_u - Q_L}{R_p\, Q_L}\right)\left(\frac{Q_L R_p}{Q_u}\right)\right]^2 = \left[\frac{Q_u - Q_L}{Q_u}\right]^2 = \left[1 - \left(\frac{Q_L}{Q_u}\right)\right]^2 = \eta$$

where η is defined as the Power Transfer Efficiency of one, real-life, lossy, LC section of an impedance matching network.

The efficiency of a passive network (i.e. its Insertion "Gain") is a number less than unity, and when expressed in decibels, it is a negative number. However, since passive network gain is more commonly described as "Insertion Loss" and as a positive number, it is more convenient to invert the mathematical expression for efficiency $\Longrightarrow 1/\eta$ and therefore obtain Insertion Loss (=I.L.) in non-negative dB:

$$I.L._{dB} = 10 \log_{10}\left[\frac{1}{\eta}\right] = 20 \log_{10}\left[\frac{Q_u}{Q_u - Q_L}\right]$$

Note: This loss is due only to power wasted through dissipation in real-life circuit components, which, no matter how good they are, can never quite match the theoretical, lossless ideal. It does not include additional losses, if any, due to power lost in reflection (VSWR) e.g. due to mis-tuning, or to radiation, etc.

APPENDIX II

(See Appendix I for definitions)

Let R_o be essentially infinite (generator lightly coupled.) Therefore $G_o = 0$. Thus:

$$\frac{P_{R_{L\ (loaded)}}}{P_{R_{L\ (unloaded)}}} = \left(\frac{G_p + G_{L_p}}{G_p}\right)^2$$

However, $\left(\frac{1}{G_p}\right) = R_p = \left[\left(\frac{1}{G_{cap.}}\right) + \left(\frac{1}{G_{ind.}}\right)\right]$

and $G_{cap} = (1/R_{P_{cap}})$

$$Q_{cap} = (R_{P_{cap}}/X_{C_P})$$

$$R_{P_{cap}} = Q_{cap}\,X_{C_P}$$

$$G_{ind} = (1/R_{P_{ind}})$$

$$Q_{ind} = (R_{P_{ind}}/X_{L_P})$$

$$R_{P_{ind}} = Q_{ind}\,X_{L_P}$$

$$(1/R_P) = 1/(Q_{cap}X_{C_P}) + 1/(Q_{ind}X_{L_P})$$

but at resonance $X_{C_P} = X_{L_P}$,

thus:

$$G_P = \left(\frac{1}{R_P}\right) = \left[\left(\frac{1}{Q_{cap}}\right) + \left(\frac{1}{Q_{ind}}\right)\right]\left(\frac{1}{X_{C_P}}\right)$$

and:

$$\left(\frac{X_{C_P}}{R_P}\right) = \left(\frac{1}{Q_u}\right) = \left[\left(\frac{1}{Q_{cap}}\right) + \left(\frac{1}{Q_{ind}}\right)\right] = \left(\frac{X_{L_P}}{R_P}\right)$$

APPENDIX III

In Figures 3 and 4 it was assumed that the available capacitor Q was greater than that of the inductor. However, from Table II of Handbook Section 2-18, it may be seen that the general tendency of capacitor-to-inductor unloaded Q ratios is to decline the closer one gets to the transistor input. It is conceivable that, at the transistor itself, especially at very high power levels (where the capacitance value is large and its Q thus low), the ratio in some cases might become less than unity (e.g. the inductive path might be short, wide, and relatively low in resistance.) In such a situation, the capacitor and not the inductor would become the predominating loss element. Where this happens, the words "capacitor" and "inductor" in Figures 3 and 4 should be mentally interchanged.

APPENDIX IV

Figure 4 of this article was originally drawn as shown below in Figure A. It was plotted using the equations and method shown below:

$$\left(\frac{1}{Q_u}\right) = \left(\frac{1}{Q_{cap}}\right) + \left(\frac{1}{Q_{ind}}\right) = \left(\left[\frac{Q_{ind}}{Q_{ind}}\right]\frac{}{Q_{cap}}\right) + \left(\left[\frac{Q_{cap}}{Q_{cap}}\right]\frac{}{Q_{ind}}\right) = \left(\frac{Q_{ind} + Q_{cap}}{Q_{cap}\quad Q_{ind}}\right)$$

$$Q_u = (Q_{cap}Q_{ind}) \,/\, (Q_{ind} + Q_{cap})$$

let $y = (Q_{cap}/Q_{ind})$

$Q_{cap} = y\, Q_{ind}$, and substituting for Q_{cap}:

$$\text{Thus } Q_u = \left[\frac{(yQ_{ind})\quad Q_{ind}}{Q_{ind} + yQ_{ind}}\right] = \frac{y\quad (Q_{ind})^2}{(1 + y)Q_{ind}}$$

$$Q_u = \left(\frac{y}{1 + y}\right) Q_{ind}$$

$$\left(\frac{Q_u}{Q_{ind}}\right) = \left(\frac{y}{1 + y}\right) \left\{\text{that is: a function of } \left(\frac{Q_{cap}}{Q_{ind}}\right)\right\}$$

Ratios of (Q_{cap}/Q_{ind}) from 1.0 to 100. were inserted, and the resulting (Q_u/Q_{ind}) ratios tabulated. To determine the percentage by which Q_u differs from Q_{ind} due to the losses in real-life capacitors, the following method was used:

$$(\%\text{ degradation}) = (100\%) - (Q_u/Q_{ind})\,(100.\%)$$

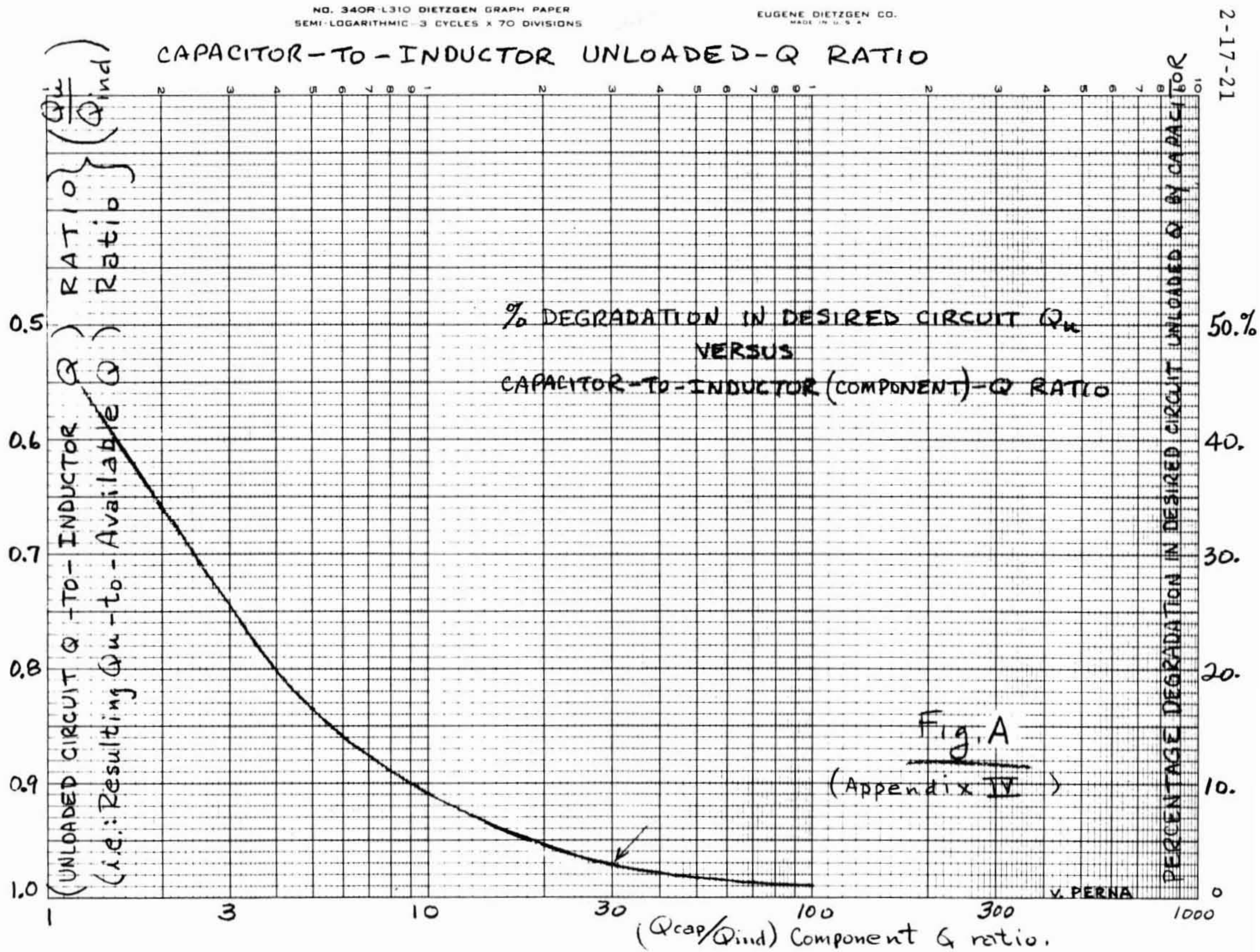
The advantage of the format of Figure A was the display of the level of the unloaded circuit Q attained versus the level of the Q available in the inductor, while simultaneously showing the percentage degradation caused by the presence of the capacitor.

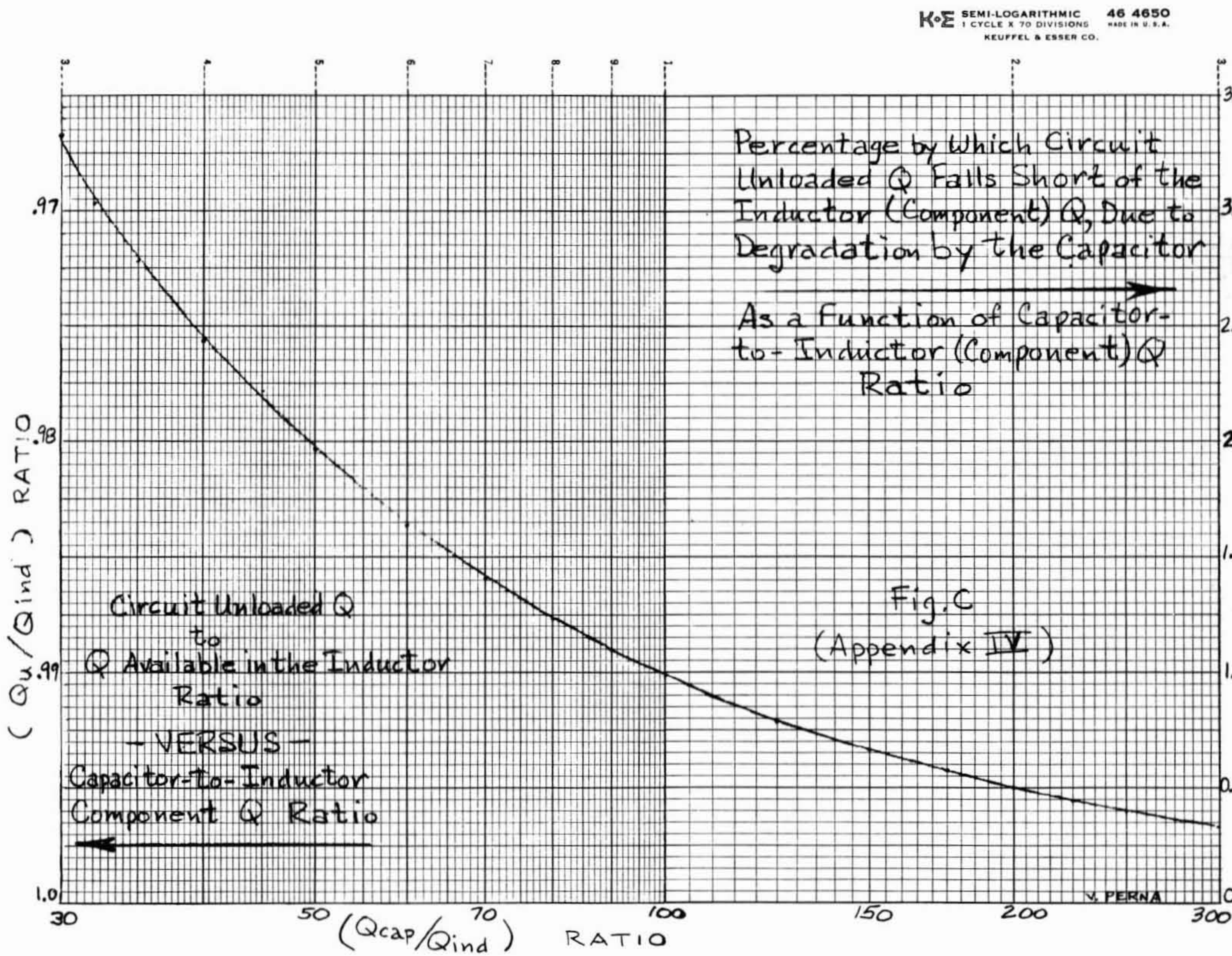
Additionally, since some low power circuits near the 50. Ohm level do not demand ultra-high Q capacitors, having the full (Qcap/Qind) scope of Figure A available seemed useful.

However, from Figure A, we can see that to obtain a circuit unloaded Q that is 97.% of that available in the inductor, we would need a (Qcap/Qind) ratio of roughly 30. Unfortunately, in this range, a small change in this ratio results in a noticeable change in network dissipation loss. Since the slope of Figure A is too compressed in this particular area for good accuracy, it was replotted in Figure 4 to gain this advantage and still give equal clarity over the whole range. Unfortunately, in the process, we lost the useful left hand scale.

For RF power amplifiers, however, the region to the right of the arrow in Figure A involved an area of more reasonable circuit losses. For this reason, and since the left hand ordinate of Figure A was so convenient for calculations, it seemed worthwhile to accept the inconvenience of restricted scope, so Figure B was developed to show this useful portion. For those willing to accept this restriction, Figure B should speed calculations.

This paper was delivered at the 15th Midwest Symposium on Circuit Theory at the University of Missouri-Rolla, May 5, 1972.


NO. 340R-L310 DIETZGEN GRAPH PAPER
SEMI-LOGARITHMIC - 3 CYCLES x 70 DIVISIONS
EUGENE DIETZGEN CO.
MADE IN U.S.A.
CAPACITOR-TO-INDUCTOR UNLOADED-Q RATIO
% DEGRADATION IN DESIRED CIRCUIT Q_u
VERSUS
CAPACITOR-TO-INDUCTOR (COMPONENT)-Q RATIO
Fig. A
(Appendix IV)
V. PERNA
(UNLOADED CIRCUIT Q -TO- INDUCTOR Q) RATIO
(i.e.: Resulting Q_u-to-Available Q) Ratio
$\left(\frac{Q_u}{Q_{ind}}\right)$
PERCENTAGE DEGRADATION IN DESIRED CIRCUIT UNLOADED Q BY CAPACITOR
0.5
0.6
0.7
0.8
0.9
1.0
50.%
40.
30.
20.
10.
0
1
3
10
30
100
300
1000
(Q_{cap}/Q_{ind}) Component Q ratio.



CAPACITOR Q FROM NETWORK EFFICIENCY

RAPID CALCULATION OF MATCHING-NETWORK, DISSIPATION-LOSS

By Vincent F. Perna
Vice President
Microwave Engineering
American Technical Ceramics

Early in the development of vacuum tube voltage amplifiers, it was determined that the voltage transfer efficiency (η) of a tank circuit[1,2,3] was given by:

$$E_{ff}(\text{voltage}) = \eta = [1 - (Q_L/Q_u)]$$

where Q_L = loaded circuit Q,

and Q_u = unloaded circuit Q.

Today, by contrast, we are concerned with high-current, low-impedance RF, power transistors and the power-transfer efficiency of their interstage networks. Unfortunately, many engineers are unaware of the historical origins of the preceding equation and erroneously try to apply it in power-amplifier calculations.

The correct relationship for power transfer efficiency[4,5,6,8] is:

$$E_{ff}(\text{power}) = (\eta) = [1 - (Q_L/Q_u)]^2$$

More convenient to use is its inverse expression in terms of loss (in dB):

$$\text{Loss (dB)} = 10 \log_{10} [1/\eta] = 20 \log_{10} [Q_u/(Q_u - Q_L)]$$

which is shown graphically in Figure 1.

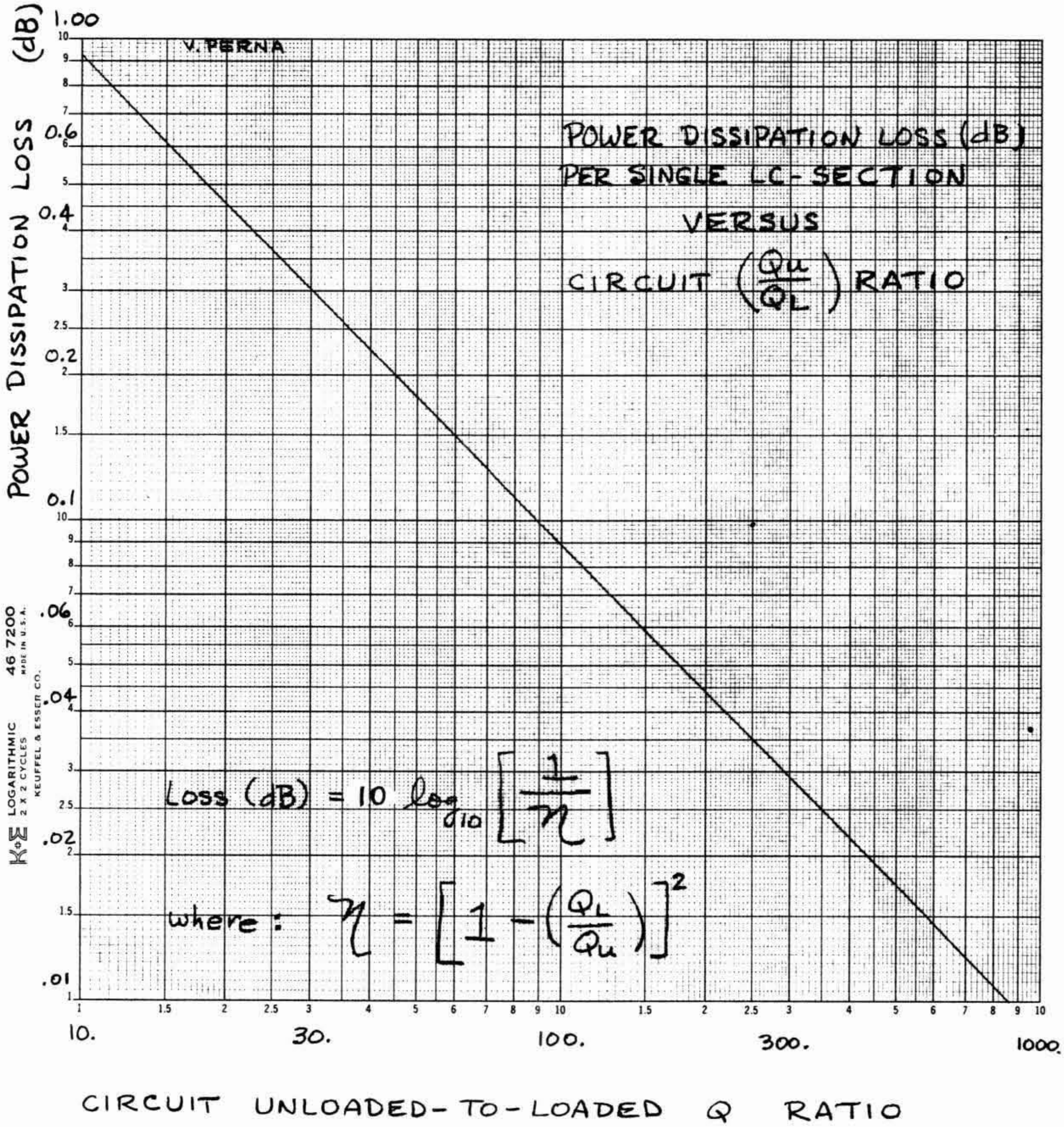

V. PERNA
POWER DISSIPATION LOSS (dB)
PER SINGLE LC-SECTION
VERSUS
CIRCUIT $\left(\frac{Q_u}{Q_L}\right)$ RATIO
$\text{Loss (dB)} = 10 \log_{10}\left[\frac{1}{\eta}\right]$
where: $\eta = \left[1 - \left(\frac{Q_L}{Q_u}\right)\right]^2$
POWER DISSIPATION LOSS (dB)
1.00
0.6
0.4
0.2
0.1
.06
.04
.02
.01
10.
30.
100.
300.
1000.
CIRCUIT UNLOADED-TO-LOADED Q RATIO
K+E LOGARITHMIC 2 X 2 CYCLES KEUFFEL & ESSER CO.
46 7200 MADE IN U.S.A.

FIGURE 1

As a practical application of this information, let us determine the overall 3-section, dissipation loss of wideband, impedance-matching network. The component Q's and loaded-circuit Q's for one such network at a specific frequency are noted in Figure 2:

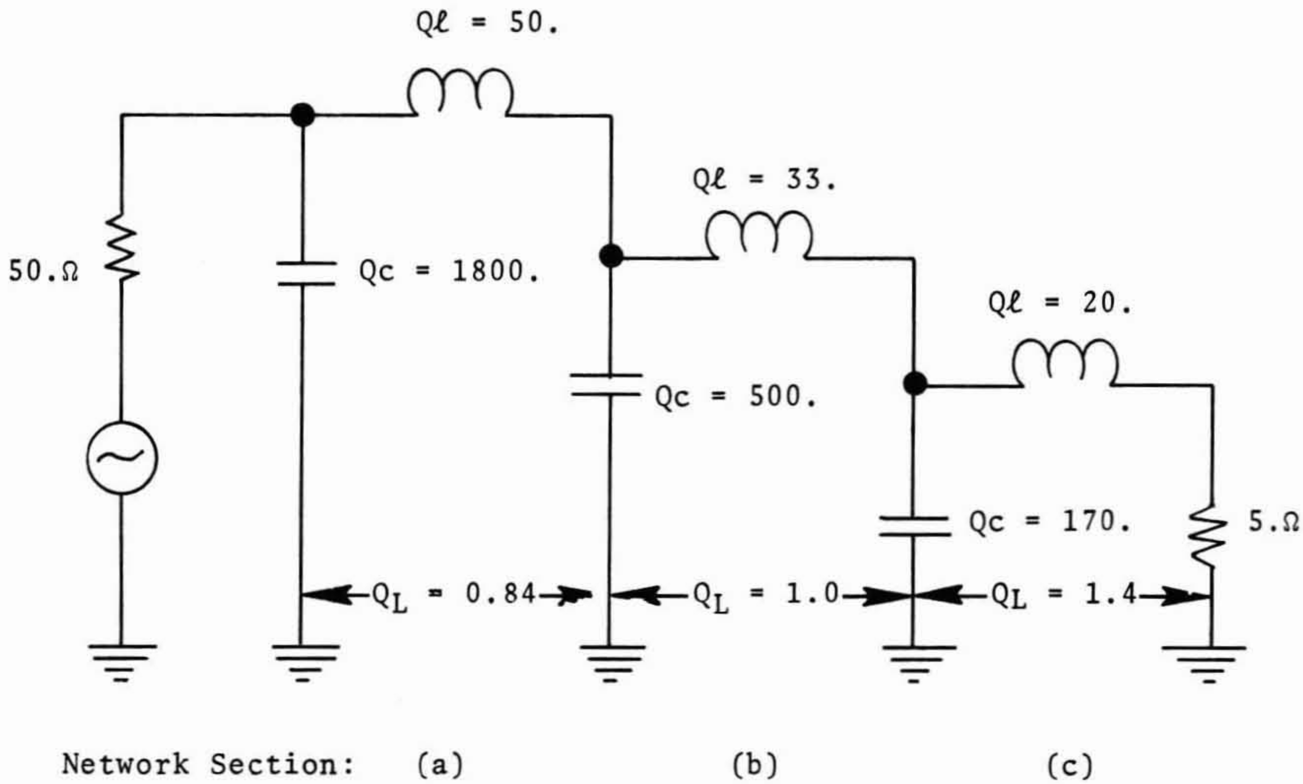


FIGURE 2

Using the equation: $(1/Q_u) = (1/Q_{cap}) + (1/Q_{ind})$ we may determine the circuit unloaded Q from the unloaded Q's of its components.

For ease of calculation, this equation is graphically displayed in Figure 3, and a step-by-step tabulation of the overall network loss at any particular point in the frequency spectrum is shown in Table 1.

The method involves the following steps for each of the several LC sections of the overall network:

(1) From the given component Q's at the frequency of interest, determine the ratio: (Q capacitor/ Q inductor).

(2) From step (1) above and the inductor Q (given in Figure 2), Figure 3 may be used to determine the unloaded Q of that section of the circuit (= Q_u).

(3) From step (2) and the loaded* circuit Q of the LC section (= Q_L) calculate (Q_u/Q_L).

(4) With the results of step (3), use Figure 1 to determine the loss in dB for that LC-section.

(5) Add up the individual LC-section dissipation-losses to get the total network loss at the frequency of interest.

---

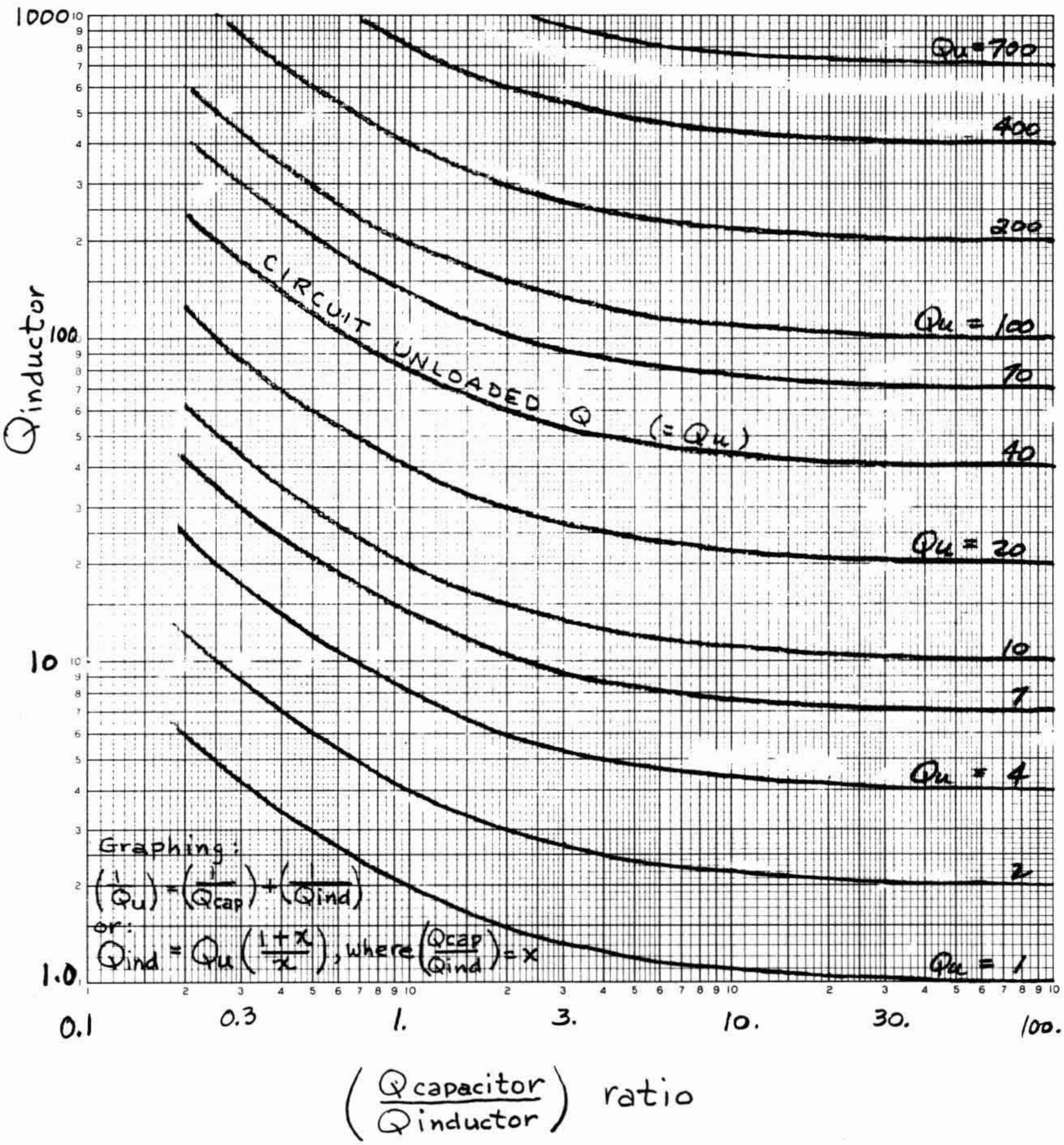
*(Q_L is usually given by the requirements of the circuit design technique. Where this data is not available, since the network may have been designed previously by someone else, Q_L may be calculated from the L/C ratio as follows:

All resistances transformed to a single series element:	$Q_L = (1/R_S)\sqrt{L/C}$
All resistances transformed to a single parallel element:	$Q_L = R_P\sqrt{C/L}$

Note: the R_S and R_P *here* are not just that of the components alone, but *include* the influence of the load resistance.

Circuit unloaded Q (= Qu) as a function of:

(a) Inductor Q
and (b) Capacitor-to-Inductor Q-ratio



Full Logarithmic, 3 × 3 Cycles

Figure 3(a)

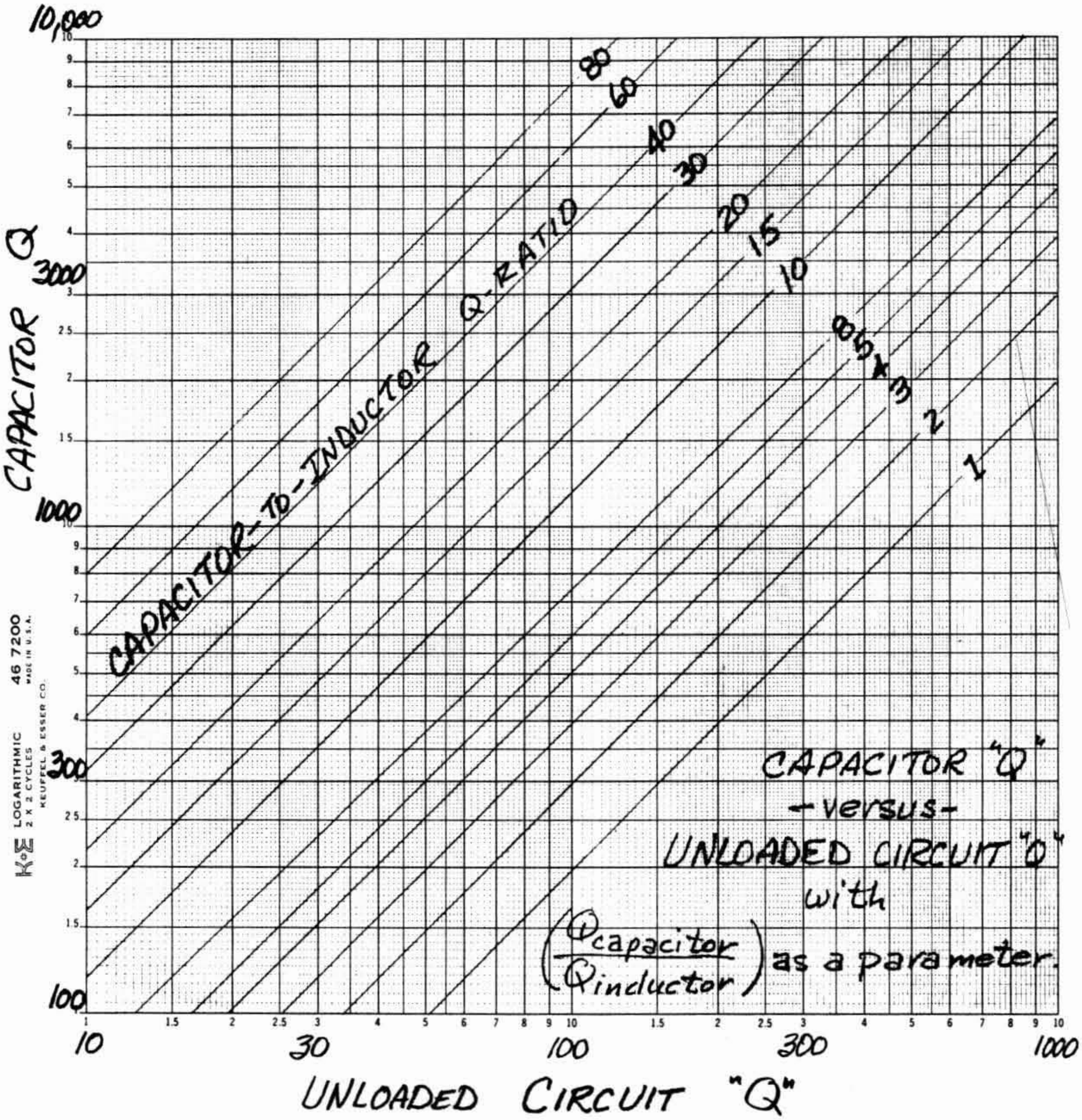

10,000
3000
1000
300
100
CAPACITOR Q
CAPACITOR-TO-INDUCTOR Q-RATIO
80
60
40
30
20
15
10
5
3
2
1
CAPACITOR "Q"
-versus-
UNLOADED CIRCUIT "Q"
with
(Qcapacitor / Qinductor) as a parameter.
10
30
100
300
1000
UNLOADED CIRCUIT "Q"
46 7200
LOGARITHMIC 2 X 2 CYCLES
KEUFFEL & ESSER CO.

FIG. 3(b)

TABLE I

CALCULATION OF CIRCUIT LOSS		NETWORK SECTION		
STEP	TO DETERMINE	(a)	(b)	(c)
(1)	(Q_{cap}/Q_{ind})	$\frac{1800.}{50.} = 36.$	$\frac{500.}{33.} = 15.$	$\frac{170.}{20.} = 8.5$
(2)	Circuit Q_u (from Fig. 3)	48.	31.	18.
(3)	(Q_u/Q_L)	$\frac{48.}{0.84} = 57.2$	$\frac{31.}{1.0} = 31.$	$\frac{18.}{1.4} = 12.9$
(4)	Loss (dB) (from Fig. 1)	0.16 dB	0.29 dB	0.71 dB

(5) From this we find that our total interstage network loss is:

(a) 0.16

(b) 0.29

(c) 0.71

1.16 dB

This is twice the amount that would have been expected from (erroneously) employing the voltage efficiency equation. Thus, the squared term in the power efficiency equation is a cause of the surprising discrepancy, experienced in the lab, between predicted and actual circuit losses.

Figure 3 is useful for a rough "eyeball" estimate. More accurate, and perhaps easier to use, is Figure 4. The steps to be followed in this case are:

(A) From the component Q's given at the frequency of interest, determine the ratio: (Q_{cap}/Q_{ind}).

(B) From the (Q_{cap}/Q_{ind}) value of step (A), and assuming that the inductor Q is the maximum that the unloaded Q of the circuit could hope to attain, use Figure 4 to determine the percentage degradation below the Q_{ind} value that the circuit will experience due to the presence of the capacitor.

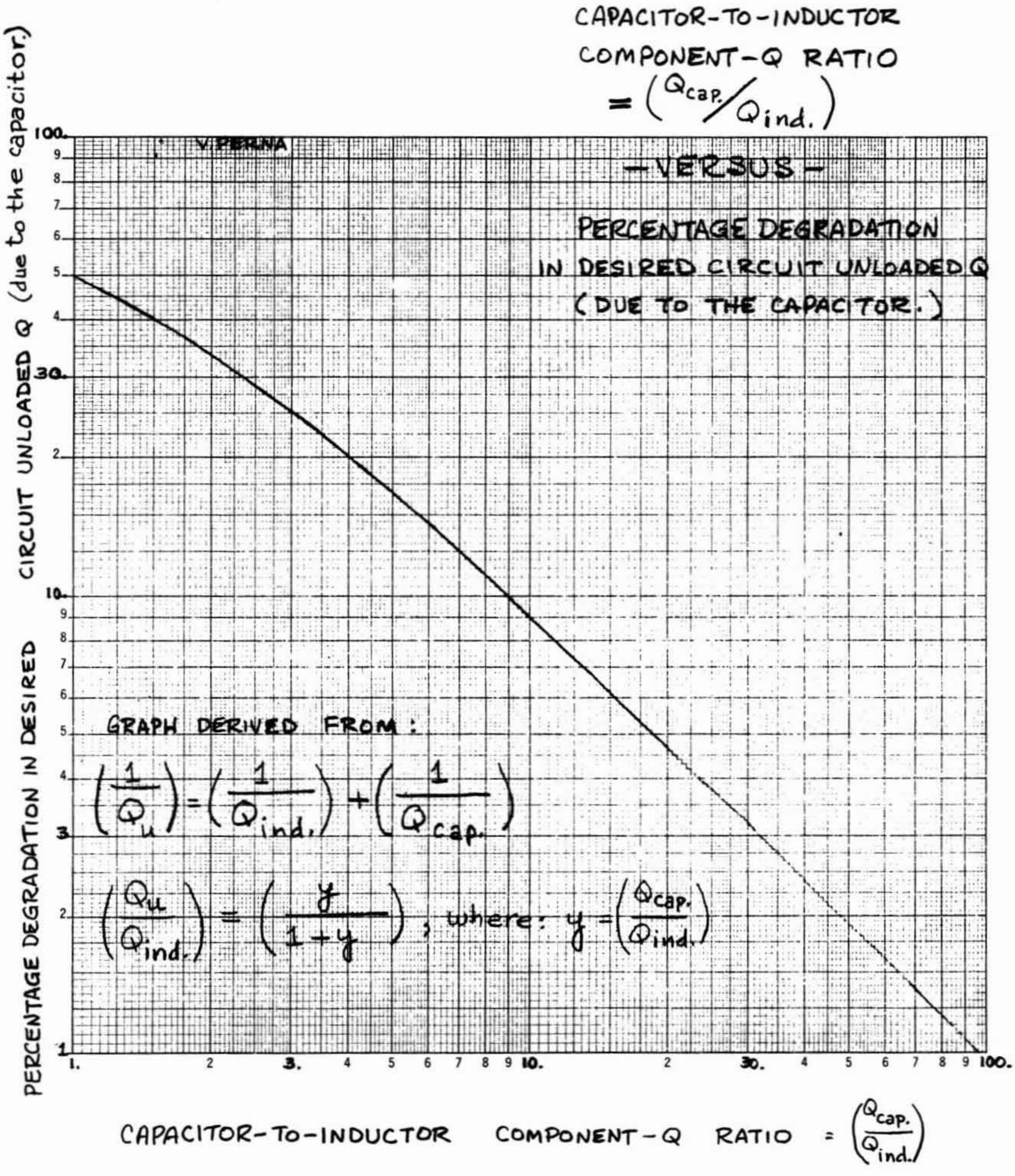
(C) Determine what the actual magnitude of the ΔQ in step (B) is.

(D) Subtract the results of step (C) from the value of Q_{ind}. This is the *unloaded* Q of the LC circuit ($=Q_u$).

(E) Using the result of step (D), and the *loaded* Circuit Q given in Figure 2, calculate the unloaded-to-loaded circuit Q ratio: (Q_u/Q_L).

(F) Using the results of step (E) and Figure 1, determine the loss per LC-section in dB.

(G) Sum the results of step (F) for each network section to obtain the total loss for the entire network.



Logarithmic, 2 x 2 Cycles

Figure 4

This process is detailed in Table II:

TABLE II

CALCULATION OF CIRCUIT LOSS		NETWORK SECTION		
STEP	TO DETERMINE	(a)	(b)	(c)
(A)	Capacitor-to-Inductor Q ratio $= (Q_{cap}/Q_{ind})$	$\left(\frac{1800.}{50.}\right) = 36$	$\left(\frac{500}{33.}\right) = 15.$	$\left(\frac{170.}{20.}\right) = 8.5$
(B)	Percentage Degradation below Q_{ind} of Q_u due to Q_{cap} (= ΔQ_{in}%)	2.65%	6.1%	10.05%
(C)	Magnitude of ΔQ (using Q_{ind} in Fig. 2)	50. X .026 = 1.3	33. X .061 = 2.01	20 X .100 = 2.0
(D)	$Q_u = Q_{ind} - \Delta Q$	50.0 -1.3 = 48.7	33.0 -2.0 = 31.0	20.0 -2.0 = 18.0
(E)	(Q_u/Q_L) = Circuit Unloaded-to-Loaded Q Ratio	$\left(\frac{48.7}{0.84}\right) = 58.$	$\left(\frac{31.0}{1.0}\right) = 31.0$	$\left(\frac{18.0}{1.4}\right) = 12.9$
(F)	Loss (dB) per LC-section from Fig. 1	0.13 dB	0.29 dB	0.71 dB

A comparison with Table I shows that, for all practical purposes, the results are essentially the same. One decided advantage of using Figure 4, however, may be ease of reading and a greater assurance of accuracy.

BIBLIOGRAPHY

1. Radio Engineer's Handbook, by F.M. Terman, McGraw-Hill, 1943, page 450.

2. Microwave Transmission Design Data, by T. Moreno, Sperry Gyroscope Co., Inc., Publication #23-80, May, 1944, page 205.

3. Very High Frequency Techniques, Harvard Radio Research Lab., Boston Technical Publishers, 1965, page 614, and 745.

4. Reference Data for Radio Engineers, Fourth Edition, ITT Corp., page 579.

5. Transistor Circuit Analysis and Design, by F.C. Fitchen, Van Nostrand, 1960, pages 249-250.

6. "How To Calculate Required Capacitor Q from Desired Matching Network Efficiency or Loss", by V.F. Perna, American Technical Ceramics, (RF Capacitor Handbook Section 2-17).

7. Radiotron Designer's Handbook, edited by F. Langford-Smith, published by Radio Corporation of America, Fourth Edition, 1953, page 411.

8. Modern Transistor Electronics Analysis and Design, by F.K. Manasse, J.A. Ekiss, and C.R. Gray, Prentice Hall, 1967, pages 144-150

9. Communication Engineering, by W.L. Everitt, McGraw-Hill, 2nd Edition, 1937, pages 260-262.

10. Radio Instruments and Measurements, National Bureau of Standards Circular #C7, U.S. Dept. of Commerce, Government Printing Office, January 1, 1937 (corrected 1924 reprint), pages 90-95.

11. "Resonance: An experimental demonstration of series and parallel resonant circuits for Radio training classes", by T.J. Rehfisch, Electronic Engineering ((British)), July, 1944, pages 76-82.

12. "Theory and Application of Resistance Tuning", by C. Brunetti and E. Weiss, Proc. IRE, June, 1941, pages 333-344.

13. "Antenna-Matching Network Efficiency", by R.L. Tanner, Electronics, Vol. 26, November, 1953, pages 142, 143.

HEAT CONCENTRATION VERSUS CAPACITOR POWER LOSS

When a physically small capacitor serves as the transmission path for large amounts of RF current, there is invariably a non-negligible amount of RF power lost through I^2R heating. The *amount* of loss is dependent upon the quality factor of the capacitor. This, however, is influenced by the temperature of the capacitor. This temperature can arise either due to: (a) the ambient temperature outside the unit in which it is located, (b) proximity (inside that unit) to a heat producing component, and/or (c) its own internally generated heat. These factors individually or combined degrade the Dissipation Factor of the capacitor and cause increased dissipation over and above what might be expected at room temperature. A generalized equation describing this situation is:

$$T_{cap} = f\ (T_A) + f\ (T_{DISS} + T_Q) + f\ (T_{LOC.}) - f\ (CR)$$

Where,

T_{cap} = Temperature of capacitor

T_A = Ambient temperature of outside air surrounding the equipment

T_E = General air temperature inside equipment cabinet or module (due to additional heat input from direct sunlight, etc.)

T_{DISS} = Capacitor Temperature due to its own I^2R power dissipation (a function of Watts-per-unit-volume).

T_{LOC} = Temperature contribution to capacitor due to local heat intensity of other sources (e.g. transistors).

T_Q = Temperature increase due to degradation of Q by heat (= ΔT_{DISS})

CR	=	Cooling rate (heat removal provided by micro-strip lines and plastic substrate touching the capacitor)
f	=	(symbolizes:) "A function of"
Q	=	Capacitor Q = (1/DISSIPATION FACTOR)

As noted earlier, we expect a certain amount of power loss in a capacitor. What we may not be aware of however, is the magnitude of the heat level attained in the capacitor due to the combined influences of the above described sources.

Table I below relates the percentage of RF power dissipated as heat to the dissipation loss in decibels.

TABLE I

Power Loss:

1.0 dB dissipation loss	$\geq$	20% of the through-put power left behind as heat
0.1 dB dissipation loss	$\geq$	2.% of the through-put power left behind as heat
0.01 dB dissipation loss	$\geq$	0.2% of the through-put power left behind as heat

Let us assume that we have a situation in which the last case in Table I applies, and we wish to put 10 Watts of RF through a capacitor that is slightly less than a 50 mil cube in dimension. We then want to determine what the watts-per-cubic-inch of RF power dissipated is. We may proceed in the following manner:

0.2 % of 10. Watts = (.002 X 10.) W = .020 Watts = 20. Milliwatts

Capacitor Volume

.050 in. X .050 in. X .035 in. = Length X Width X Thickness

$= 5. \times 10^{-2} \times 5. \times 10^{-2} \times 3.5 \times 10^{-2} = 87.5 \times 10^{-6}$ (cubic Inches)

<u>Watts-per-Unit-Volume (i.e.: temperature)</u>:

$$\left(\frac{.020 \text{ Watts}}{87.5 \times 10^{-6} \text{ cubic inches}}\right) = \left(\frac{1.14 \times .020 \text{ Watts}}{100. \times 10^{-6} \text{ cubic inches}}\right)$$

$$= \left(\frac{.023 \text{ Watts}}{1 \times 10^{-4} \text{ cubic inches}}\right) = 2.3 \times 10^{-2} \times 10^{4} \text{ (Watts/Cubic Inches)}$$

$$= 230. \text{ Watts-per-cubic inch (of heat intensity).}$$

Thus, we can see if such a small capacitor were dissipating only .02% of the 10 Watts of power input and not cooling itself in any manner that the temperature which this tiny unit would attain would be hotter than most commonly used soldering irons.

Borrowing a quote (Ref. 1) from the RF Power Transistor field, we see that such heat concentrations are not at all unusual in RF power amplifiers:

"Aluminum metal undergoes electromigration at a rapid rate at current densities exceeding 1×10^5 A/cm^2 and film temperatures above 125°C. It is interesting to note that the concept of electromigration of metals is not too difficult to accept if one considers the power density dissipated in aluminum (volume resistivity = 2.7 μ Ω cm at 120°C) operating at a current density of 1×10^6 A/cm^2 is 2.7×10^6 W/cm^3."

The only reason that the solder joints are not melted is that the metallic lands on which the capacitor is seated, plus any other components to which they are attached (such as the substrate), carry the heat off at a rate greater than that at which it is generated.

If, however, the rate of heat generation exceeds the rate of cooling, there will be an almost instantaneous rise in the temperature of the capacitor due to thermal run away. This occurs, since, once the Dissipation Factor begins to climb significantly, it then results in an increased generation of heat (internally), which results in a rise in the Dissipation Factor, and so on ad infinitum. This will result in the destruction of the capacitor if the power from the generator is great enough (plus possibly the destruction of other circuit components as well.)

The way to avoid such problems is to use capacitors whose Dissipation Factor is so low that the internally generated heat under the anticipated conditions of operation will never be great enough to cause thermal run away.

Even where thermal runaway *is* barely avoided, there is the problem of circuit detuning due to significant capacitance shift at temperatures beyond some manufacturers' rated limits. These temperatures can easily occur when capacitors are carrying significant RF current and mounted close to a heat source inside equipment which is subjected to an outside ambient temperature of 125°C.

Thus far, we have discussed only the case of high RF current. High RF voltage can also be a source of destruction of capacitors. According to Paschen's Law(2), a void or an air bubble in a dielectric between two electrodes can have its entrapped gas become ionized. Once ionized, the gas becomes an electrical conductor and all electrical conductors have resistance which generates heat during the transmission of RF energy. If the gas becomes sufficiently hot, it can become a plasma which absorbs large amounts of RF energy directly, thus increasing the heat intensity. The plasma then melts the surrounding ceramic, further increasing the gas pressure. The process continues until the walls of the capacitor melt and blow out (or the capacitor body in total explodes). The remains, when minutely examined, may resemble a miniature volcano which has blown its top.

One recent advertisement by a capacitor manufacturer claimed that he had "fewer holes" (voids) within his capacitors than his competition (Brands X, Y and Z). By comparison, ATC capacitors have no holes. Consequently, they may be used at power levels up to 15 KW (pulsed), and over 1 KW CW at the 50 Ohm level without destruction. If there were voids, there would be volcanos.

Another unpleasant feature of voids is they allow internal migration of metal which can cause a short circuit. In fact, if there is sufficient porosity, this may provide penetration paths which will allow entrance and entrapment of moisture and contaminants which degrade performance.

From these facts we can see that poor construction methods or a low quality factor dielectric can result in intense heating, and thus poor circuit performance.

Use of ATC 100 Series low loss microwave porcelain capacitors avoids these problems.

BIBLIOGRAPHY

1.) "RF Power Transistor Metallization Failure", by James R. Black, IEEE Transactions on Electron Devices, Vol. ED-17, No. 9, September, 1970, pages 800-803.

2.) High Voltage Technology, Oxford University Press, 1968, page 45.

3.) Hybrid Microelectronics Review, Vol. III, No.6, June, 1972, "Rule 3: Power---The Forgotten Parameter" (address: P.O. Box 11685, Philadelphia, Pennsylvania 19116).

PRE-DETERMINATION OF DC-TO-RF CONVERSION EFFICIENCY

by Vincent F. Perna, Jr.
Vice President
Microwave Engineering

Accurately predicting the DC-to-RF conversion efficiency of proposed equipment has largely remained an art whose success is related to the extent of the engineer's RF background.

This can cause serious problems when, in times of keen competition, non-RF companies commit themselves to a contract which will require extensive RF circuitry. Many painful months later, they may discover that one of their unexpected stumbling blocks was the perhaps innocent-looking restriction on available DC current.

This article hopes to minimize such problems by injecting more science into this art.

It is not commonly realized that there are several historical reasons for the present difficulty.

A good number of years ago the term "collector efficiency":

$$\text{Eff. (collector)} = \left(\frac{\text{RF out}}{\text{DC in}}\right) \qquad (1)$$

came into common use and was generally satisfactory at lower frequencies for comparing transistors as oscillators or where amplifier stage-gains were high.

In more recent times, where frequencies of operation and power levels have risen, the more accurate expression "overall efficiency":

$$\text{Eff. (overall)} = \left(\frac{\text{RF out}}{\text{DC in + RF in}}\right) \qquad (2)$$

for comparing transistors has also been employed. It has the advantage that it helps account for the fact that transistor gains at UHF and above are much lower and there are often significant amounts of forward-signal feed-through.

The problem is that, by this time, the designation: "collector efficiency" had become a generally employed standard of comparison on data sheets, and thus had become trapped there.

In addition, in high-power UHF amplifiers, circuit losses had become quite an important consideration due to the rapid fall-off of passive-component Q with increasing frequency. Furthermore, *serious drawbacks of both the "collector"-and "overall-" efficiency concepts are that they include circuit losses but do not indicate either their magnitude, the circuit configuration, the bandwidth, or the extent of unwanted additional losses in both the semiconductor and external circuit from unspecified harmonic currents flowing.*

As a result, the actual semiconductor efficiency is significantly higher than either of these figures of merit would lead one to assume.

To clarify matters, it is helpful to separate the problem into its two main elements:

- the power-transfer efficiency of the interstage matching networks

  and
- the efficiency of the semiconductor itself.

CIRCUIT EFFICIENCY

Attempts at calculating circuit losses have encountered the difficulty that modern texts on the subject are often ambiguous or even in seeming disagreement on the mathematical expression for circuit efficiency.

One commonly listed equation is:

$$\eta = \left[1 - \left(\frac{Q_L}{Q_u}\right)\right] \tag{3}$$

What is not widely realized by readers, however, is that this is the expression for the singly-loaded power transfer-efficiency of a tank circuit, developed in the early days of low frequency "Radio", when vacuum tube output resistances were high compared to load resistances.

As might be expected, this equation entered the literature of the period, along with a discussion of its significance. Apparently, however, as more and more useful knowledge of radio circuits was compiled, the handbooks became overpacked with data and had to be condensed. In many cases, the explanatory text was crowded out.

The result is today, 50 years later, we have some amazingly complete assemblages of equations, but sometimes a less than complete idea of how to employ them effectively.

Further problems are caused by the fact that today we are less concerned with high-impedance vacuum tube amplifiers and are increasingly more involved with high-current, low-impedance, RF power transistors.

Unfortunately, many engineers are unaware of the origins of equation (3) and erroneously try to apply it where both source and load resistances are low.

Although there are several aspects of such calculations which could benefit by clarification, this present article will attempt to tackle only one: the need to develop an accurate means of pre-determining the DC-to-RF conversion efficiency of an RF amplifier---regardless of circuit configuration.

After extensive study of the literature in comparison with known power transmission-versus-efficiency relations on a slide rule, it became clear that the proper expression for the doubly-loaded power-transfer efficiency of a single, LC matching-section was actually given by:

$$\text{Eff. (power)} = [\eta] = \left[1 - \left(\frac{Q_L}{Q_u}\right)\right]^2 \qquad (4)$$

where Q_L = Doubly-Loaded-circuit Q,
and Q_u = Unloaded-Circuit Q.

The Q_u of the network has been known for decades to be directly related to the component Q's of the circuit elements, and that to attain the highest circuit efficiency it was advisable to employ components with the lowest equivalent-series-resistance. More recently, in the low-impedance, high-current portions of modern high-power RF semiconductor circuits, attention to this fact has become imperative.

Using this as a basis of decision, once the circuit elements have been selected, the resulting total interstage matching network loss may be determined in steps, beginning at one end of the network and working through, totalling the losses of individual LC sections.

First, the unloaded Q=(Q_u) of each individual matching section is given by Equation (5):

$$\left(\frac{1}{Q_u}\right) = \left(\frac{1}{Q_{cap.}}\right) + \left(\frac{1}{Q_{ind.}}\right) \qquad (5)$$

To avoid time-wasting calculations, Figure 1 [derived from a rearrangement of equation (5)] has been provided to determine directly the unloaded circuit Q from: the known capacitor Q and the capacitor-to-inductor Q ratio. For example, if the capacitor Q = 1000 and the capacitor-to-inductor Q-ratio = 10, then from Figure 1 the unloaded circuit Q = 86.

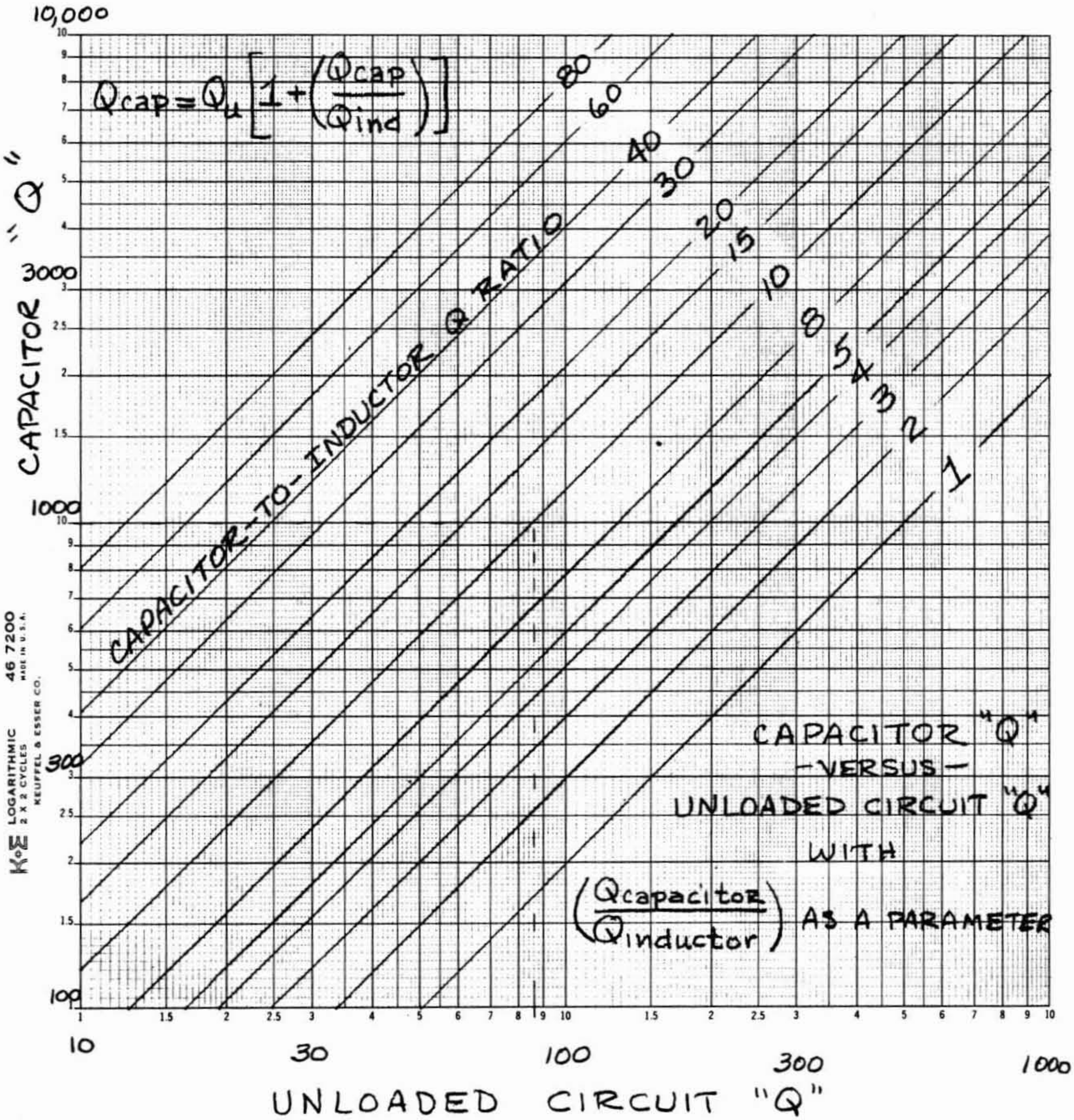

Qcap = Qu [1 + (Qcap / Qind)]
10,000
3000
1000
300
100
CAPACITOR "Q"
CAPACITOR-TO-INDUCTOR Q RATIO
80
60
40
30
20
15
10
8
5
4
3
2
1
CAPACITOR "Q"
-VERSUS-
UNLOADED CIRCUIT "Q"
WITH
(Qcapacitor / Qinductor) AS A PARAMETER
10
30
100
300
1000
UNLOADED CIRCUIT "Q"
K+E LOGARITHMIC 2 X 2 CYCLES KEUFFEL & ESSER CO.
46 7200 MADE IN U.S.A.

Figure 1

Next, the bandwidth of each section of the matching network will provide the loaded-circuit Q ($=Q_L$) for that section from Equation (6):

$$Q_L = \left(\frac{f_o}{BW}\right) \text{ , where:} \qquad (6)$$

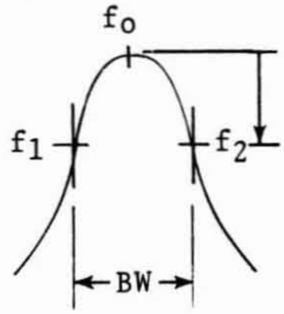


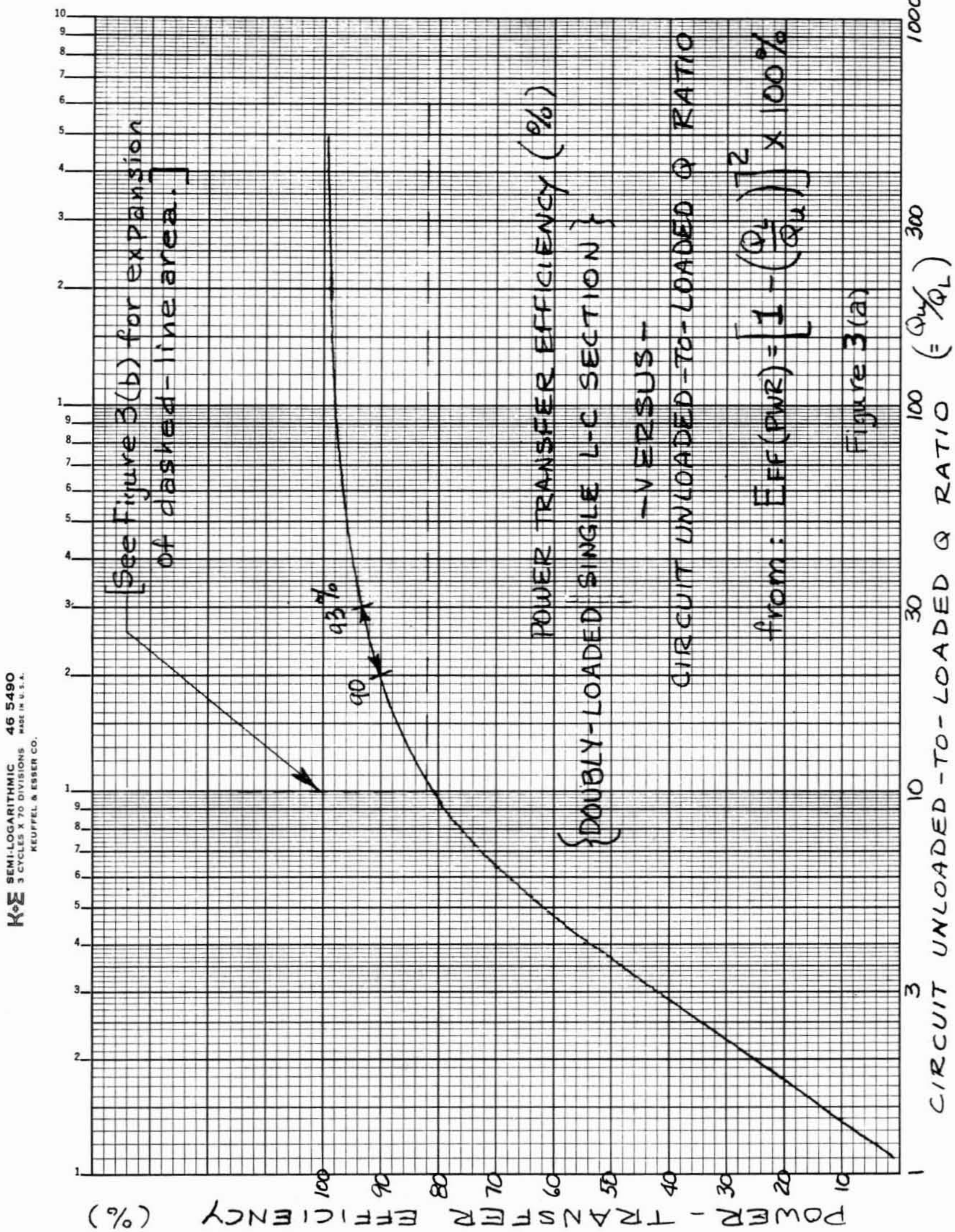
FIGURE 2

From the unloaded-to-loaded Q ratio ($=Q_u/Q_L$) for each LC section, the designer may then determine the power-transfer efficiency for that section of his interstage network (from Equation 4 or Figure 3.)

These individual, single-section efficiency-factors, when multiplied together, will give the overall power-transfer efficiency of each complete interstage-network.

SEMICONDUCTOR EFFICIENCY

One method of fairly closely determining the semiconductor efficiency might be to (a) match the input and output of the device using multi-stub tuners (Figure 4a), (b) calculate the resulting "overall" efficiency, (c) remove the semiconductor and match up to 50. Ohms again from each tuner's open port using an additional tuner at each port (Figure 4b), (d) measure the Insertion Loss of each tuner pair, (e) make the rough assumption that the losses in each tuner pair are about equally divided between the tuners of the pair, and divide the measured Insertion Losses of the input and output systems by two, (f) correct the original "overall" efficiency figure by increasing the "RF in" Wattage by the amount of the input tuner loss, the "RF out" Wattage by the amount of a the stage gain times the input tuner loss, and by the output tuner loss, (g) determine the semiconductor efficiency from equation (7):



[See Figure 3(b) for expansion of dashed-line area.]
POWER TRANSFER EFFICIENCY (%)
{DOUBLY-LOADED SINGLE L-C SECTION}
-VERSUS-
CIRCUIT UNLOADED-TO-LOADED Q RATIO
from: $E_{FF(PWR)} = \left[1 - \left(\frac{Q_L}{Q_U}\right)\right]^2 \times 100\%$
93%
90
CIRCUIT UNLOADED-TO-LOADED Q RATIO $\left(= Q_U / Q_L\right)$
POWER-TRANSFER EFFICIENCY (%)
1
3
10
30
100
300
1000
10
20
30
40
50
60
70
80
90
100
K&E SEMI-LOGARITHMIC 3 CYCLES X 70 DIVISIONS KEUFFEL & ESSER CO. 46 5490 MADE IN U.S.A.

Figure 3(a)

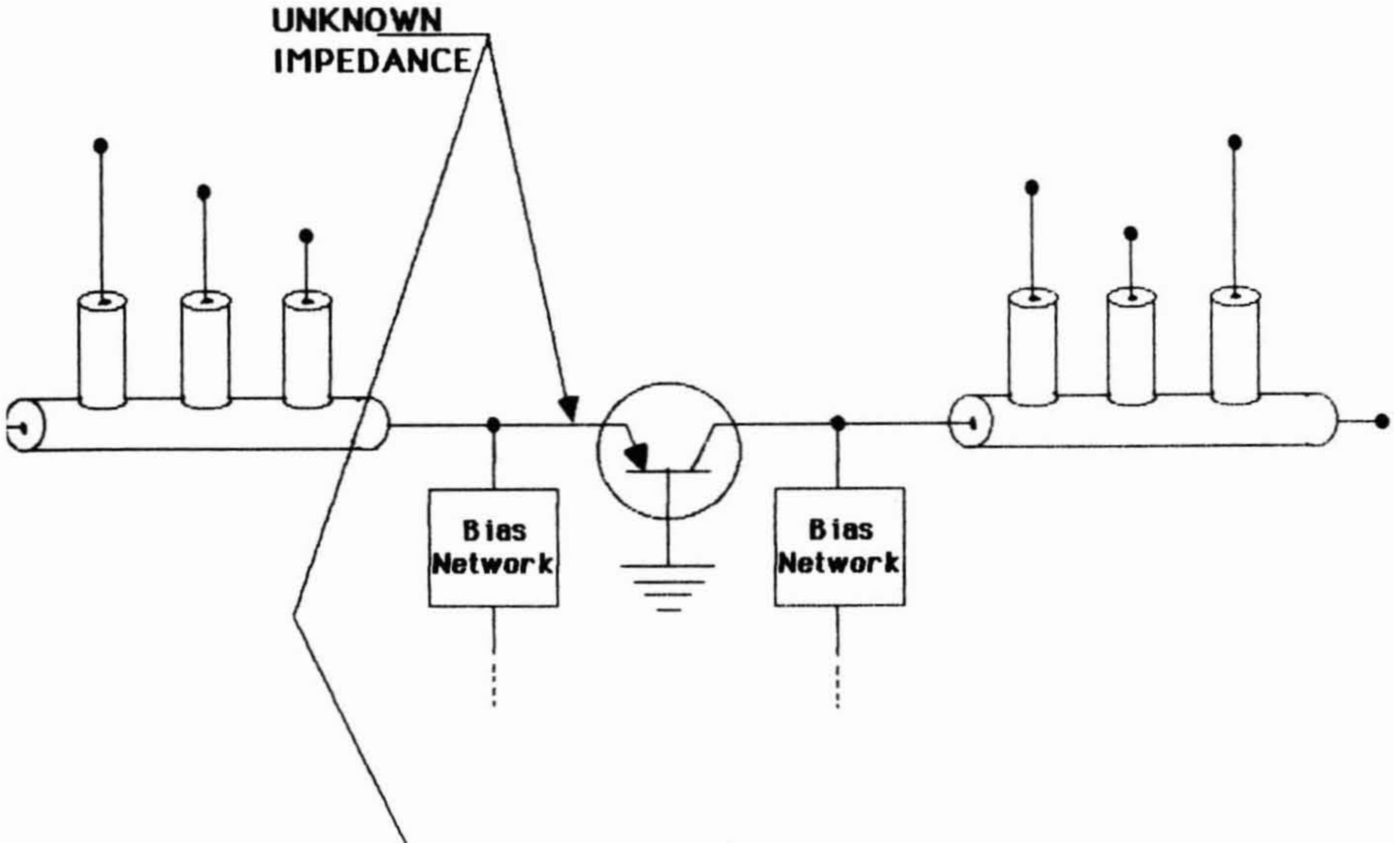


FIGURE 4 (a)

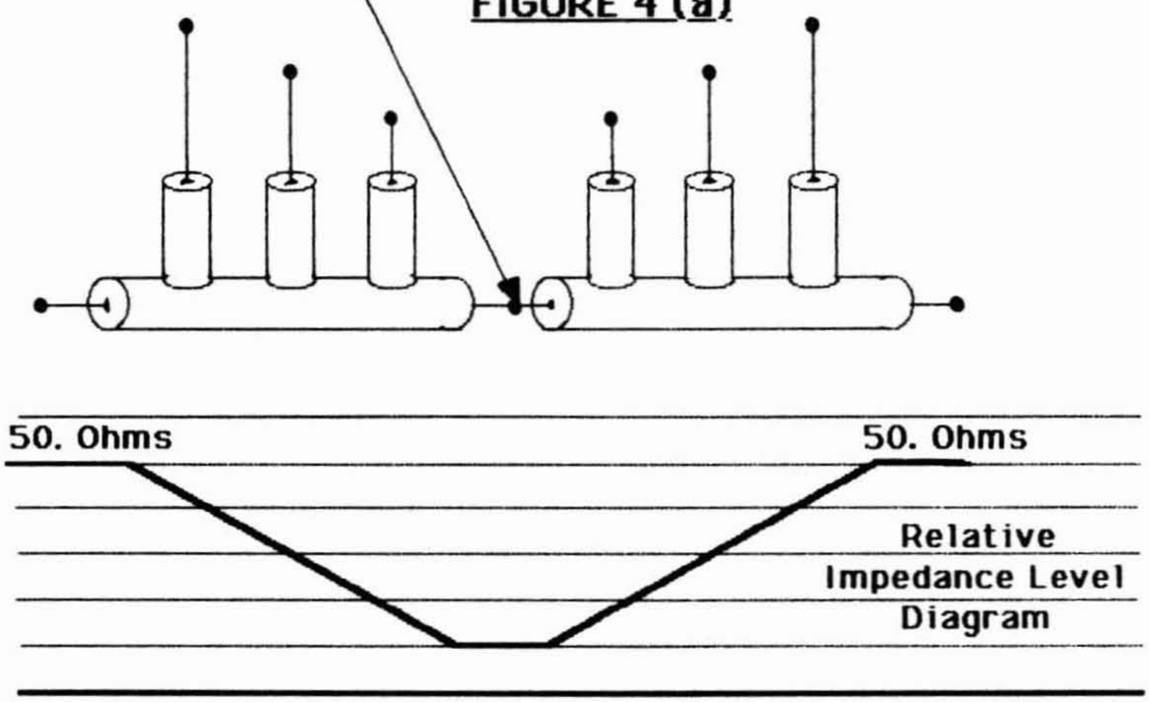


FIGURE 4 (b)

$$\text{Eff. (semiconductor)} = \left[\frac{\text{Eff ("overall", corrected)}}{\text{Eff (interstage)}} \right] \qquad (7)$$

where: the numerator is derived from step (f), and the denominator is the power-transfer-efficiency-equivalent of the sum of the individual single-tuner losses of step (e) [employ Figure 5 .]

DEVICE CHARACTERIZATION

Data on actual semiconductor device efficiency might eventually come to be expressed in some family-of-curves format such as:

$$R_{out} = \frac{(V_{cc} - V_{sat})^2}{2P_{out}}$$

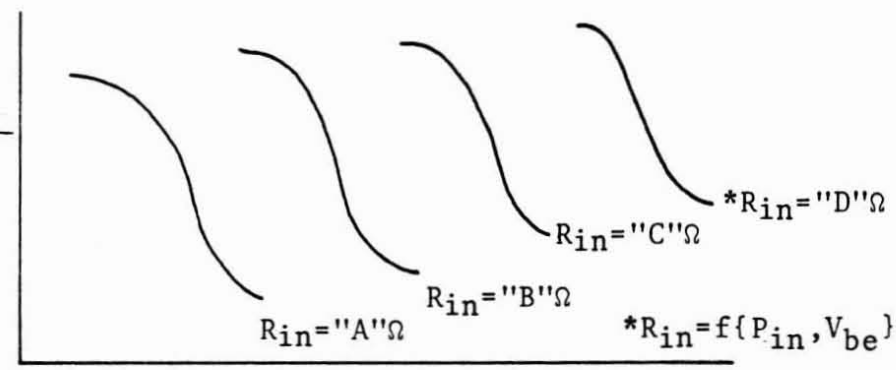


FIGURE 6

Although this presentation may require some measurement effort, it has the benefit of effectively correlating R_{in}, R_{out}, RF power level (in *and* out), and semiconductor efficiency. Perhaps more importantly, it would move the present design situation from the experimental closer to the analytical.

One weakness of the present methods of examining the immittances of transistors is the inability to accurately determine very low resistance values when using a 50. Ohm referenced measuring system. Under these circumstances, the real part of the input impedance to be measured (say, 2. Ohms) may be

POWER TRANSFER RELATIONSHIPS

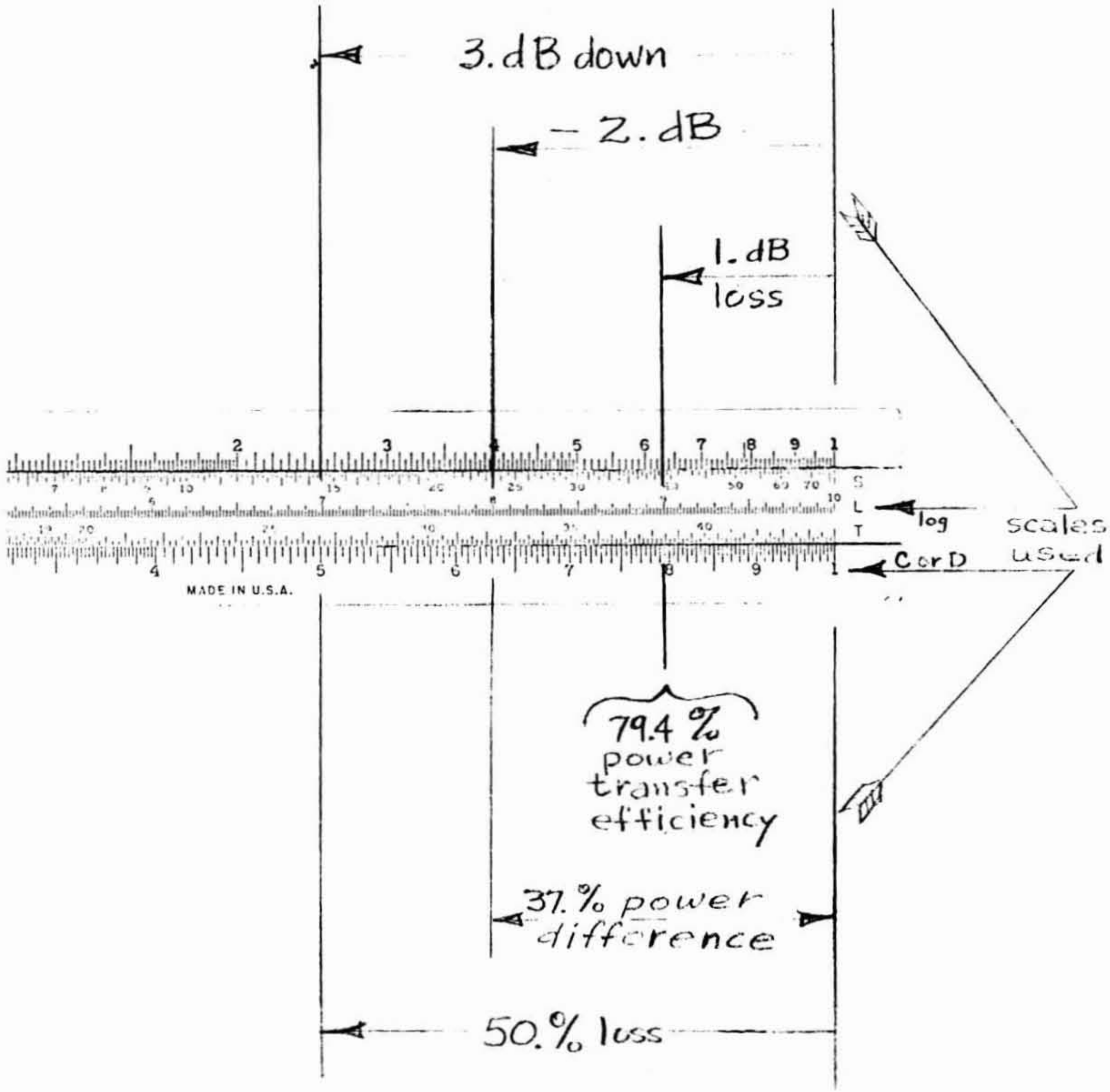


FIGURE 5

the same order of magnitude as the uncertainty of the system. This condition has discouraged many theoreticians and designers from pursuing meaningful studies.

However, if we were to use the set-up of Figure 7 , we could change the center of the Smith Chart from 50. Ohms down to 5. or 10. Ohms, greatly reducing this uncertainty, and significantly increasing the accuracy of measurement. The extent of the success of this method may be seen in references (26) and (27) in the Bibliography.

LIMITATIONS ON THE METHOD

Since R_{in}, circulating current, bandwidth, and gain are inter-related, if Figure 6 is to be a practical aid, efficiency may have to be quoted on some specific percentage-bandwidth basis.

APPLYING THE DATA

Active component (semiconductor) efficiency times the total interstage network (passive) efficiency provides the to-be-expected, amplifier-stage efficiency factor. (Subdivision into passive and active-portion efficiencies has the benefit of providing insight into which areas may require closer investigation, for example, component Q and/or transistor efficiency versus temperature.)

Finally, multiplying together each of the amplifier-stage efficiency-factors would provide the total DC-to-RF conversion efficiency for the amplifier chain.

Using this method might help industry avoid the pitfalls presently encountered when relying upon educated guesstimates alone.

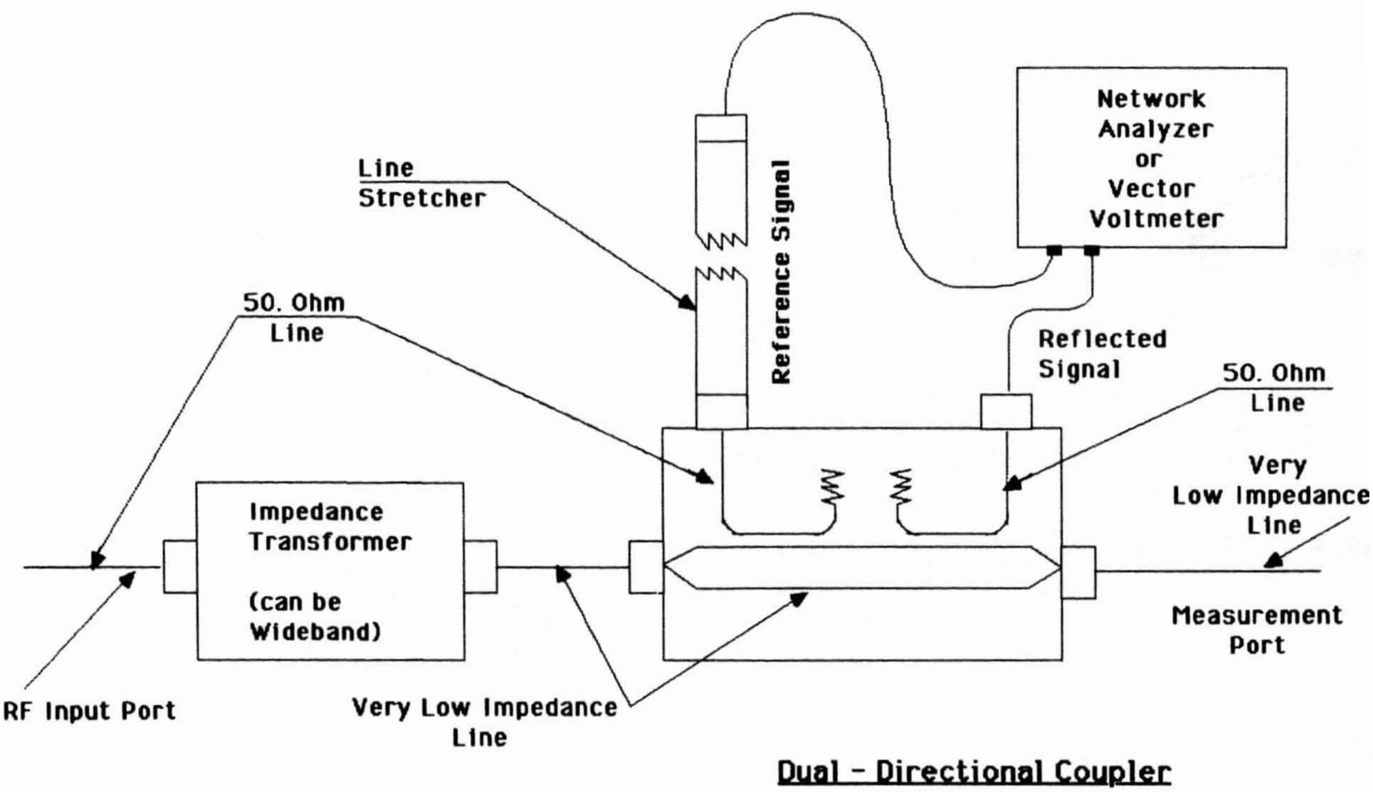


FIGURE 7.

DC-RF CONVERSION EFFICIENCY

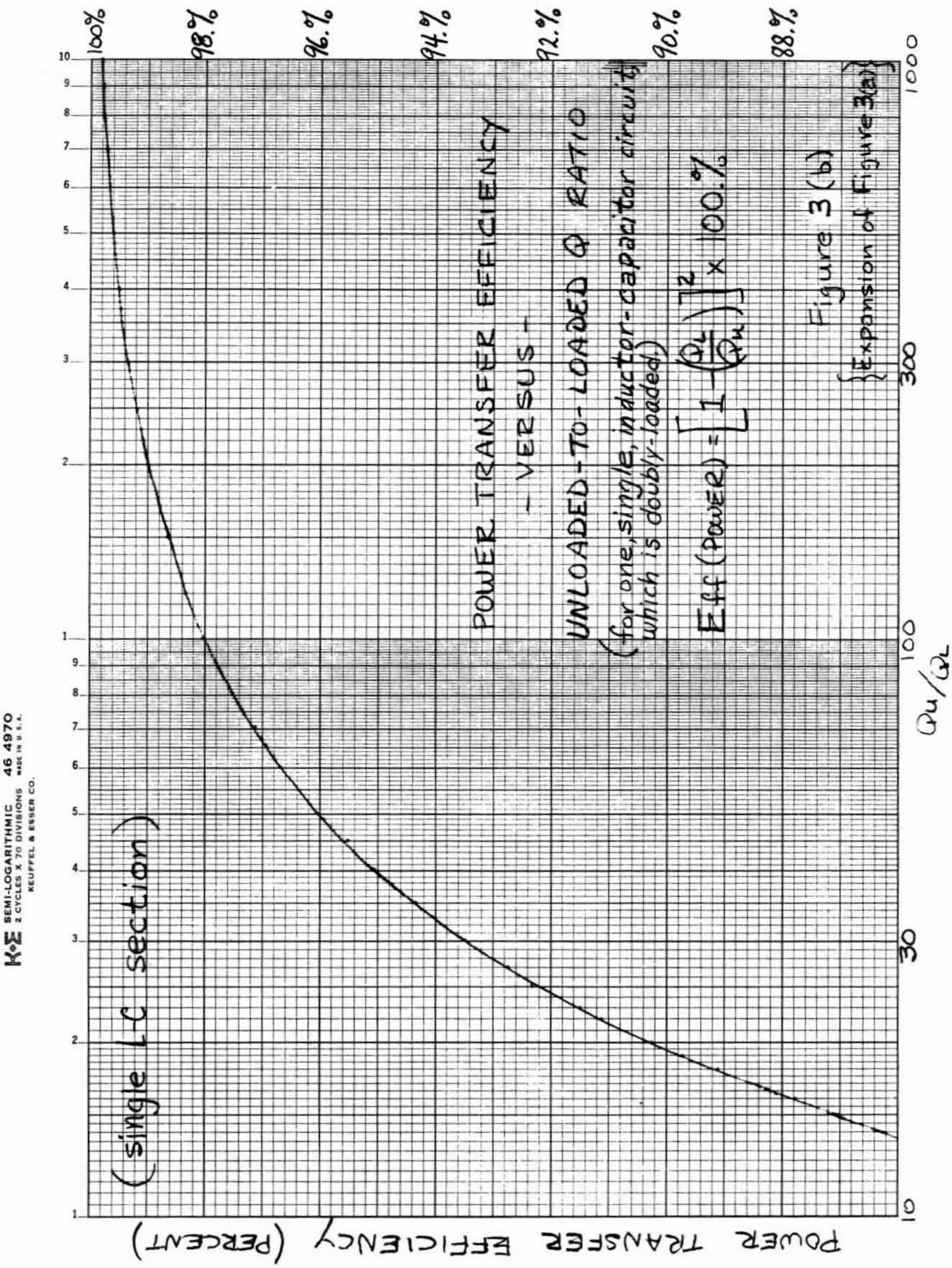


Figure 3(b)
{Exposion of Figure 3(a)}

BIBLIOGRAPHY

For references 1 through 25, see the Bibliography of Section 2-17.

(26) Technical Report ECOM-0503-P001-G777 "Wideband Stripline 10. Ohm Measurement System" by Max Richardson, Oct. 1968 Written for Sylvania under contract to Fort Monmouth, N.J.

(27) Technical Report ECOM-0503-P001-G724 "Design Considerations for Wideband High Power Solid State Amplifiers" by W.T. Hatley, Jr., S.J. Klein, G.L. Hatley, Feb., 1969.

(28) This method is applicable to more accurately determining θ_j also. See page 7 of the uncondensed version of the article by J.H. Johnson and M.J. Mallinger of CTC entitled "The RF Transistor Comes of Age" which appeared in Electronics Magazine, September 13, 1971.

DC–RF CONVERSION EFFICIENCY

QUOTABLE QUOTES ON THE BENEFITS OF ATC CAPACITORS IN "LOW NOISE FRONT ENDS"

2-21a From "MICROSTRIP AMPLIFIERS CAN BE SIMPLE", (by George D. O'Clock, Jr., Electronic Design 14, July 8, 1971, pages 66-68.)

"The frequency response is primarily limited by the gain-bandwidth product (f_t) of the transistor and the quality of the chip capacitor elements. In this case, Nippon Electric (NEC) 2N5761 transistors were used (f_t = 3.7 GHz to 4.5 GHz) along with American Technical Ceramic porcelain chip capacitor elements. The frequency response of the amplifier indicates an f_t of approximately 3.5 GHz for each transistor stage. Therefore, the frequency-response degradation caused by the capacitor chips appears to be minimal."

2-21b Quotes and Abstracts from "METHODS OF REDUCING NOISE OF JUNCTION FIELD EFFECT TRANSISTOR (JFET) AMPLIFIERS", (by H.E. Kern and J.M McKenzie, IEEE Transaction on Nuclear Science, Vol. NS-17, No. 1, February, 1970, pp. 260-268.)

"Radeka's work and our own parallel studies have shown an important noise source to be lossy dielectric materials in---capacitors,---"

"The need for high resolution or low noise, particularly in the energy range required for analytical work" ((below 200 eV)), "is apparent from these curves."

"Radeka has shown that lossy dielectric materials in critical input capacitors---contributes significantly to the noise at high frequencies" ((i.e. 100 KHz.))

"Equation (1) shows that the higher the" ((parallel, R_p, or shunt leakage)) "loss resistance of a dielectric material, the lower will be its noise contribution."

"Porcelain capacitors manufactured by American Technical Ceramics Corp. exhibit very high" ((shunt)) "loss resistance" ((= very low leakage current)). "These capacitors---can be used quite successfully as calibrating or feedback capacitors in low noise amplifier circuits."

2-21c UHF RADAR FRONT ENDS

In a UHF phased-array radar, ATC 100 low-loss microwave porcelain capacitors were used in the receiver's front-end to attain a Noise Figure of 1.4 dB. Since the semiconductor device had a NF of 1.2 dB, there was only 0.2 dB left for circuit losses from all causes. Thus, even commonly-accepted, loss sources had to be minimized here.

This accomplishment was also important *economically*, since to attain long-range capability, the only alternative was to raise transmitter power---but this would have cost kilo-bucks per kilowatt. It was less expensive to reduce the receiver's Noise Figure.

NOISE FIGURE CONTRIBUTION OF MIC MATCHING SECTIONS

By Vincent F. Perna, Jr., Vice President, Microwave Engineering
American Technical Ceramics

One school of thought on Noise Figure degradation holds that the thermal noise contribution of circuit elements is a function of the relative magnitudes of the individual resistances involved in the circuit. From this viewpoint, the base spreading resistance (and its tightness of coupling to the source resistance) in relation to the circuit gain strongly determines front-end Noise Figure. Using this approach, a good 50. Ω input impedance is impossible, and to meet the Noise Figure specification, one may have to make VSWR trade-offs and perhaps select a noise measuring instrument from the lab whose inaccuracy is known to read in the company's favor.

Another school of thought maintains that any dissipation of signals in the circuit elements of the receiver front-end contributes directly to system Noise Figure. Thus, only high-Q, low loss, components are acceptable if the combination of a decent input match, high gain, and low NF is to be attained. This latter approach may require interactive loss calculations, and to speed up the process, the following method is offered.

If we say that we have a receiver front-end which contains individual impedance matching sections of the form of Figure 1, we may desire to know the loss contribution to the Noise Figure of the system of each such section.

Figure 1 might represent the last in a chain of three such LC sections in a preselector. If we allow about 0.05 dB

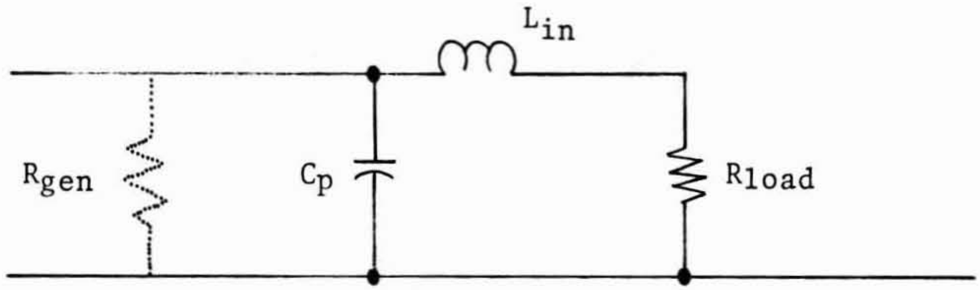


Figure 1

dissipation for soldering (and sundry other signal path losses), and approximately equal dissipative losses per LC section, the per-section signal degradation would be 0.15 dB.

Now let us assume we have a specification for a section of a receiving system which says that it must exhibit a Noise Figure of 3. dB, and our available transistor provides us 2.5 dB. To see if this is realistically attainable, before investing time and money in a breadboard model, it is instructive to calculate the matching network efficiency based upon the Q's of commercially available or attainable circuit elements and the desired bandwidth.

From Figure 2, we can see that 0.15 dB loss requires that each LC-section have an unloaded-to-loaded Q ratio ($=Q_u/Q_L$) of 60.

If we further assume an LC section bandwidth of 66.% (= a Q_L of 1.5), we then require an unloaded Q for the LC section of:

$$Q_u = 60\ Q_L = 60 \times 1.5 = 90. \tag{4}$$

From equation (5): $(1/Q_u) = (1/Q_{cap}) + (1/Q_{ind})$, (5)

we see that if the capacitor involved had an infinite Q, we would only need an inductor with a Q of 90 to attain the desired Noise Figure for that section.

In real life however, we would probably be happy to find an inductor with a Q that is consistently, say, 3.% higher in Q than our desired value. In that case, we must employ a capacitor which does not degrade the circuit Q_u more than 3.%, if that section is not to exceed a 0.15 dB Noise Figure contribution.

Since reviewing these trade-offs are often iterative procedures, we would not want to have to repeat calculations any more than necessary, therefore, Figure (2) was devised from reorganizing equation (5) into that of equation (6):

$$(Q_u/Q_{ind}) = [y/(1 + y)], \text{ where } y = (Q_{cap}/Q_{ind}) \qquad (6)$$

and plotting (6) as in figure (3): after modification as in equation (7):

$$(\%\text{ degradation}) = (100.\%) - (Q_L/Q_{ind})\ (100.\%) \qquad (7)$$

Examining Figure (3), we see that a degradation in circuit Q_u of no more than 3.% by the capacitor requires a (Q_{cap}/Q_{ind}) ratio of at least 32.

With this information, we can use equation (6) to determine the required capacitor Q. Alternatively, Figure (4) may be used for more rapid results. From this, we may see that for a Q_u of 90 and (Q_{cap}/Q_{ind}) of 32, the required Q_{cap} will be 2900.

If the capacitance values we need is 51. pF at 400. MHz, Figure 5 shows that the best available capacitor Q is 200., which is a far cry from the desired $Q_{cap} = 2900$.

We are now faced with the need to accept a higher Noise Figure contribution from such an LC-section than we had anticipated, and must reverse the calculation procedure to determine, from the available L and C, what Noise Figure we must live with.

We know now that: $(Q_{cap}/Q_{ind}) = 200/93 = 2.15$ and thus from Figure (4) that the circuit Q_u which we must live with is 62.

The required loaded circuit Q is still $Q_L = 1.5$, therefore $(Q_u/Q_L) = 62/1.5 = 41.3$.

From Figure (2), we see that the Noise Figure corresponding to this (Q_u/Q_L) ratio is 0.22 dB.

This is a NF contribution for one LC section which is 50.% greater than what we desired. If there were three such sections, we would end up with a total Noise Figure for our subsystem of:

NF = 2.5 dB + 3 X .22 = 3.16 or nearly 0.2 dB greater than what is allowed.

Consequently, for high performance MIC receiving systems, the Q of the capacitor at even low microwave frequencies can be of more importance than general belief would have led us to suspect.

Since the capacitors are usually the only area in which there is any flexibility in Q (since inductors are so low in Q to begin with), to gain an improvement in N.F. may demand an extremely large increase in capacitor Q. If this is unpallatable, some alternatives are: buying a transistor with a lower Noise Figure (= more money), accepting an increase in overall Noise Figure (= poorer system sensitivity), reducing the preselector's I.L. by widening the bandwidth and/or reducing the skirt slope (= degraded image and spurious-signal rejection, *plus* making a trade-off between lower I.L. and increased KTB noise-power feed-thru.)

In conclusion then, quoting H.T. Friis, the originator of the Noise Figure equation (1): "From a design point of view it is important to know the relationship between the Noise Figure of a whole receiver and the Noise Figures of its components since it indicates the component on which efforts for improvement are worth while."

The procedure developed in this paper now allows a simple, rapid determination of the Noise Figure contribution we can expect from the dissipation loss of available commercial capacitors and inductors, and from this determine where to concentrate efforts for improvement or trade-offs.

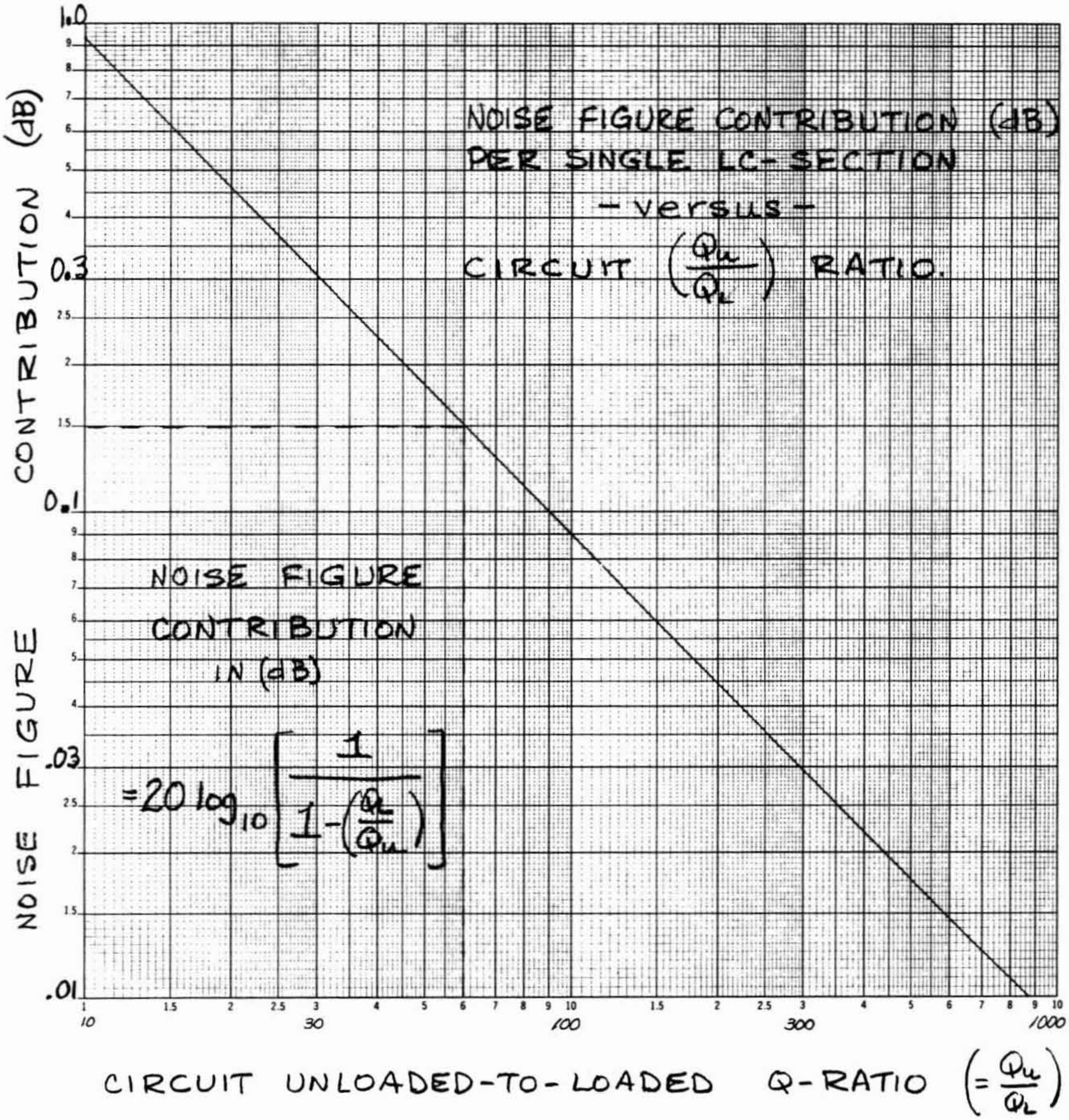

NOISE FIGURE CONTRIBUTION (dB) PER SINGLE LC-SECTION
- versus -
CIRCUIT $\left(\frac{Q_u}{Q_L}\right)$ RATIO.
NOISE FIGURE CONTRIBUTION IN (dB)
$= 20 \log_{10}\left[\frac{1}{1-\left(\frac{Q_L}{Q_u}\right)}\right]$
NOISE FIGURE CONTRIBUTION (dB)
1.0
0.3
0.1
.03
.01
10
30
100
300
1000
CIRCUIT UNLOADED-TO-LOADED Q-RATIO $\left(=\frac{Q_u}{Q_L}\right)$

Figure 2

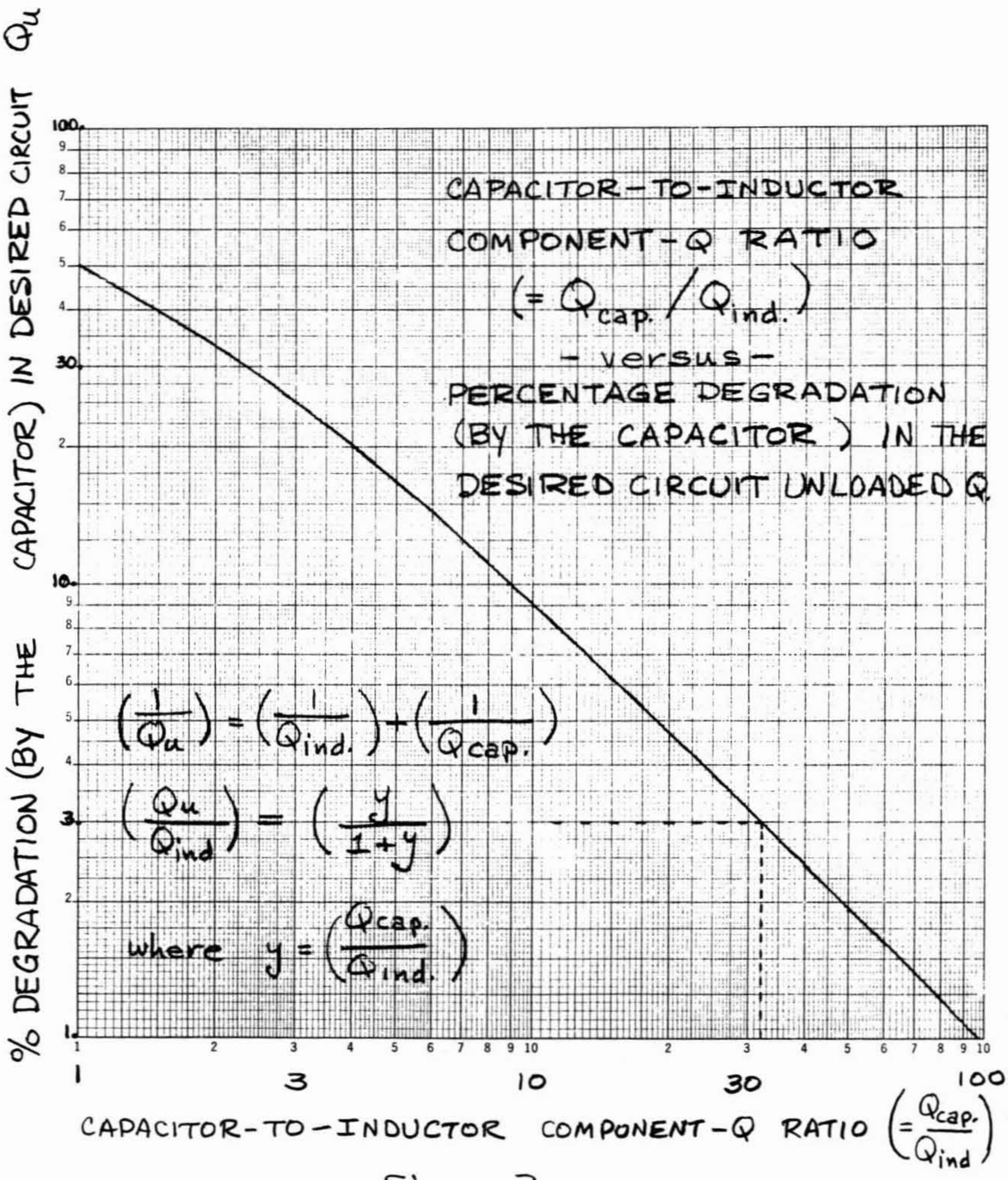
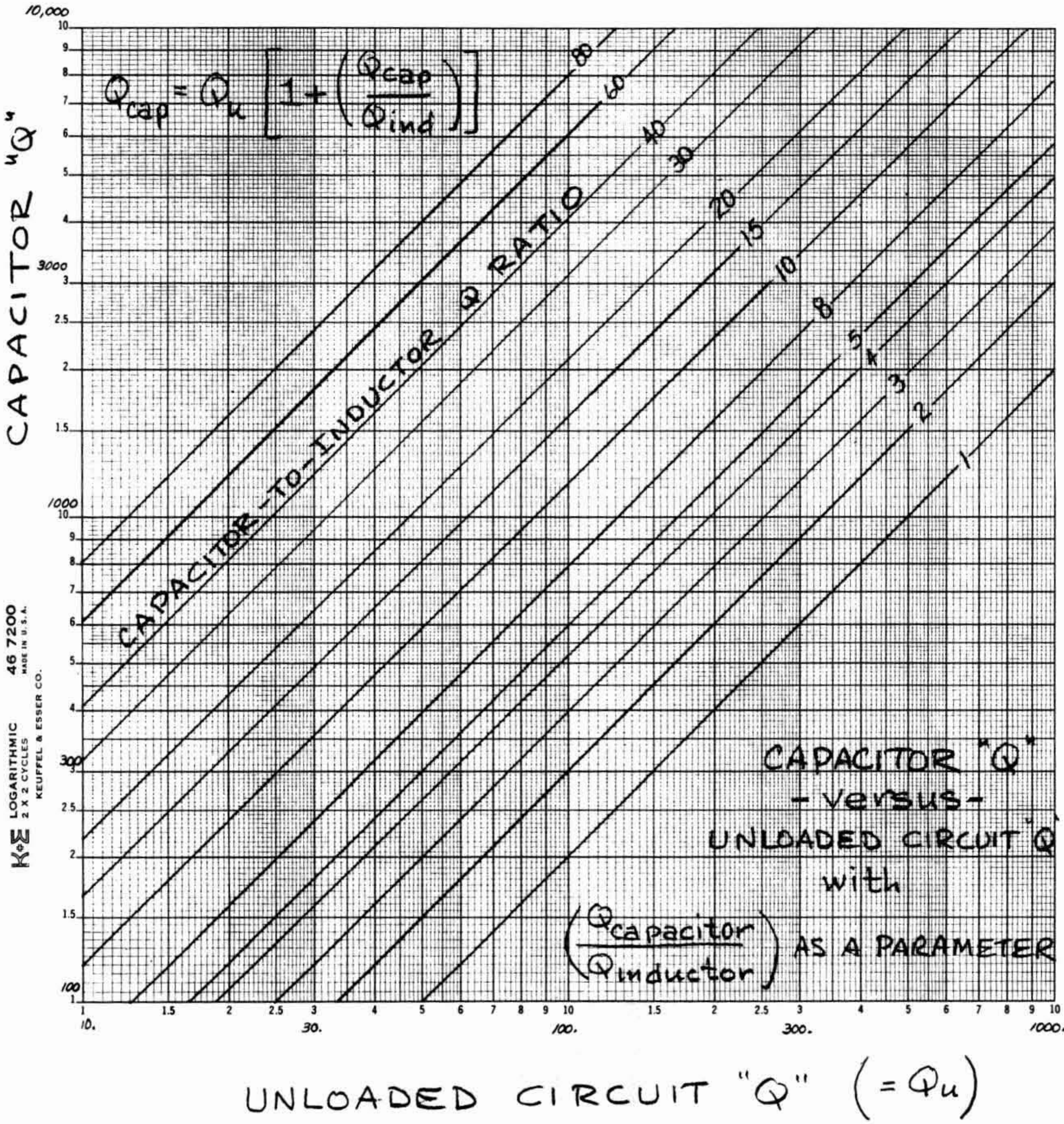


Figure 3

Logarithmic, 2 x 2 Cycles

NOISE FIGURE OF MIC



$Q_{cap} = Q_u \left[1 + \left(\frac{Q_{cap}}{Q_{ind}}\right)\right]$
CAPACITOR "Q"
CAPACITOR-TO-INDUCTOR Q RATIO
80
60
40
30
20
15
10
8
5
4
3
2
1
CAPACITOR "Q" - versus - UNLOADED CIRCUIT "Q" with $\left(\frac{Q_{capacitor}}{Q_{inductor}}\right)$ AS A PARAMETER
10,000
3000
1000
300
100
10.
30.
100.
300.
1000.
K+E LOGARITHMIC 2 X 2 CYCLES KEUFFEL & ESSER CO.
46 7200 MADE IN U.S.A.
UNLOADED CIRCUIT "Q" $(=Q_u)$

Figure 4

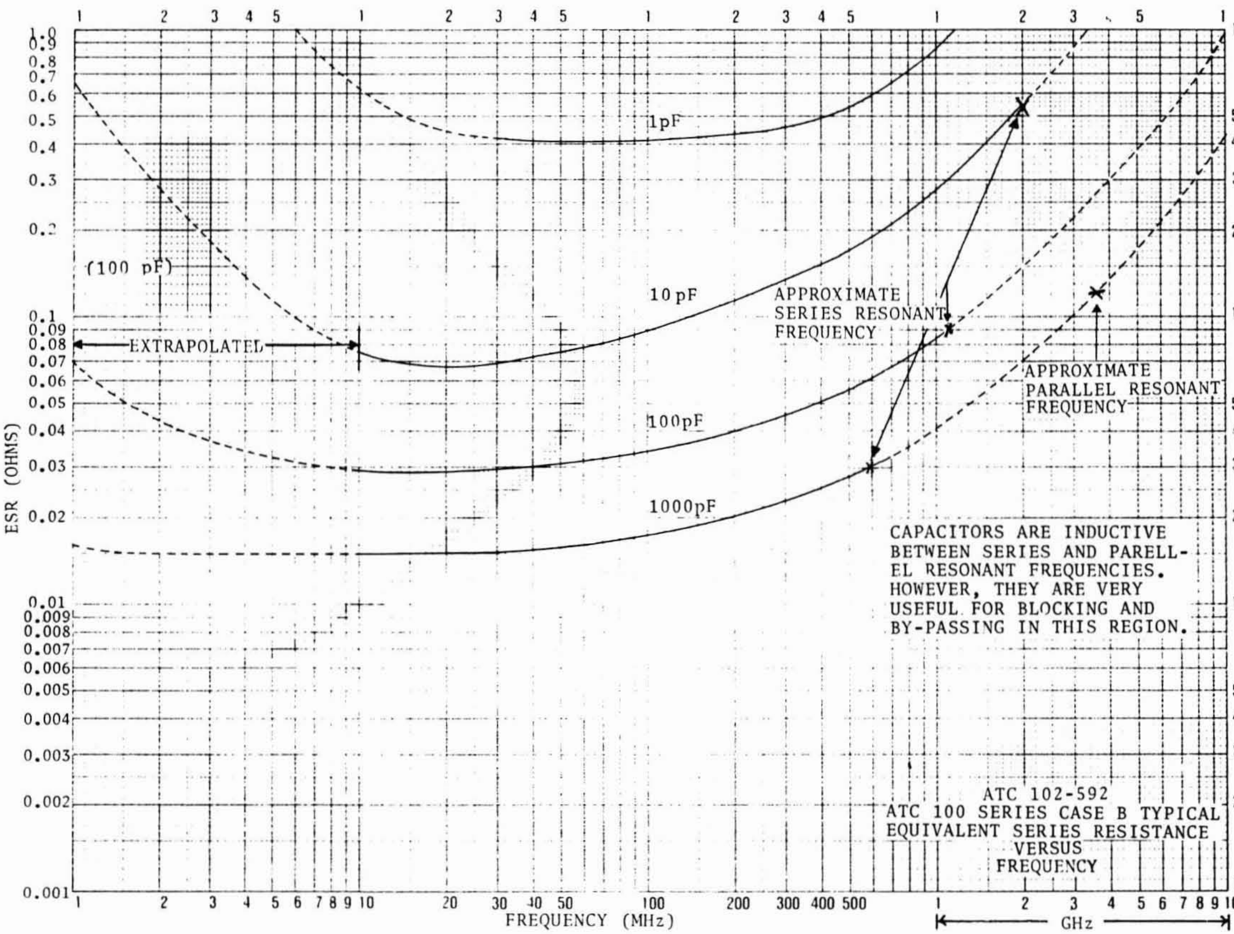

1pF
10 pF
100pF
1000pF
(100 pF)
EXTRAPOLATED
APPROXIMATE SERIES RESONANT FREQUENCY
APPROXIMATE PARALLEL RESONANT FREQUENCY
CAPACITORS ARE INDUCTIVE BETWEEN SERIES AND PARELL-EL RESONANT FREQUENCIES. HOWEVER, THEY ARE VERY USEFUL FOR BLOCKING AND BY-PASSING IN THIS REGION.
ATC 102-592
ATC 100 SERIES CASE B TYPICAL EQUIVALENT SERIES RESISTANCE VERSUS FREQUENCY
ESR (OHMS)
FREQUENCY (MHz)
GHz

FIG. 5

BIBLIOGRAPHY

1. "Noise Figures of Radio Receivers", by H.T. Friis, Proceedings of the I.R.E., July, 1944, pages 419-422.

2. "Input Circuit Noise Calculations for FM and Television Receivers", by W.J. Stolze, Communications, February, 1947.

3. "Electronic Designer's Handbook", by R.W. Landee, D.C. Davis, and A.P. Albrecht, McGraw-Hill, 1957, pages 7-19, to 7-24.

4. "Modern Transistor Electronics Analysis and Design", by F.K. Manasse, J.A. Ekiss, C.R. Gray, Prentice-Hall, 1967, pages 425, 426.

5. "Reference Data for Radio Engineers" (4th Edition) by the International Telephone and Telegraph Corporation, New York, New York, page 769.

6. Radiotron Designers Handbook (4th Edition), Edited by F. Langford-Smith, The Wireless Press, Sydney, Australia, 1953, distributed by Radio Corporation of America, Harrison, New Jersey.

7. "The Over-all Noise Factor", by H.F. Starke, Raytheon Semiconductor Engineering, File No. 134-T, Dec. 7, 1955.

8. "Noise Figure Primer", Hewlett-Packard Application Note 57 of 1/15/65.

DEFINITIONS OF SYMBOLS USED IN SECTION 2

Q_{cap} = Component (capacitor) Quality factor = $(1/D.F.) = \left(\frac{X_C}{R_s}\right) = \left(\frac{R_p}{X_c}\right)$.

D.F. = Dissipation factor

X_C = Capacitive reactance = $\left(\frac{1}{2\pi fC}\right)$

R_s = Equivalent series resistance (to alternating current of a path which cannot pass direct current.)

R_p = Equivalent parallel resistance (= resistive representation of all shunt losses involved.)

Q_{ind} = Component (inductor) Quality factor = $\frac{X_L}{R_s} = \frac{R_p}{X_L}$

X_L = Inductive reactance = $\frac{1}{2\pi fL}$

Q_u = the unloaded Q of an RLC circuit. Sometimes determined by using *extremely* light coupling, then measuring the -3 dB bandwidth, and calculating from $Q = \left(\frac{f_o}{f_2 - f_1}\right)$. Also, given at resonance by (R_p/X_C) or (R_p/X_L) .

tan δ = Loss tangent (which, in good dielectrics at high frequencies, = D.F.)

I.L. = Insertion Loss (the total combination of losses due to reflection, dissipation, radiation, etc.)

R_{eff} = the effective input resistance at band-center of a transistor plus some external tuning reactance (usually a shunt capacitor.)

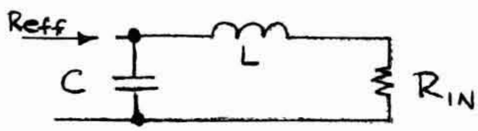
R_{in} = the input resistance of the transistor at the semiconductor chip itself.

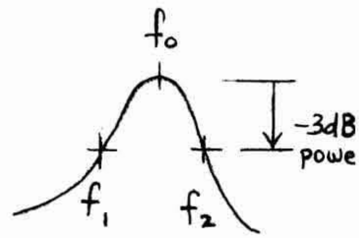
I_{in} = the RF current into R_{eff}.

I_{circ} = the RF current circulating in the network comprised of the transistor and its tuning element (e.g. a shunt capacitor.) $I_{circ} = Q_L\ I_{in}$

Q_L = Loaded circuit Q $= \left(\frac{f_o}{BW}\right)$

$$= \sqrt{\left(\frac{R_{eff}}{R_{in}}\right) - 1} = \left(\frac{R_{eff}}{X_c}\right) = \left(\frac{X_L}{R_{in}}\right)$$

for the circuit: 



BW :(Instantaneous) $= \left(f_2 - f_1\right)$ at the -3dB points.

NOTE: This is not to be confused with:

BW :(Tuning) = range over which the f_o of the circuit may be moved without seriously distorting the required $f_2 - f_1$ characteristic.

F = Noise Factor when expressed as a sum of power ratios (as in): $F_{12} = L + F_1 + \left(\frac{F_2 - 1}{G_1}\right) + ---$

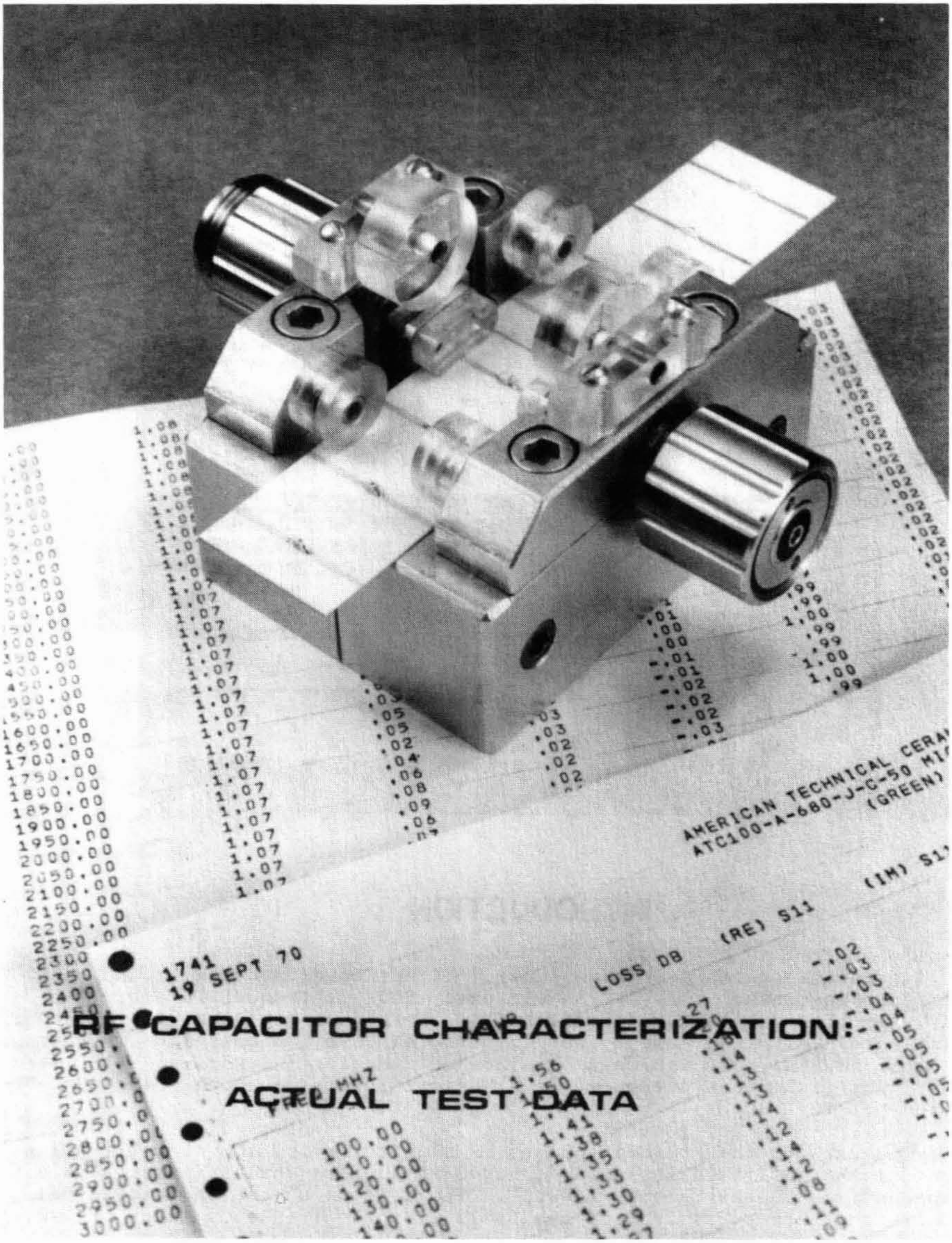
or:

N.F. = Noise Figure when the Noise Factor is expressed in terms of decibels.

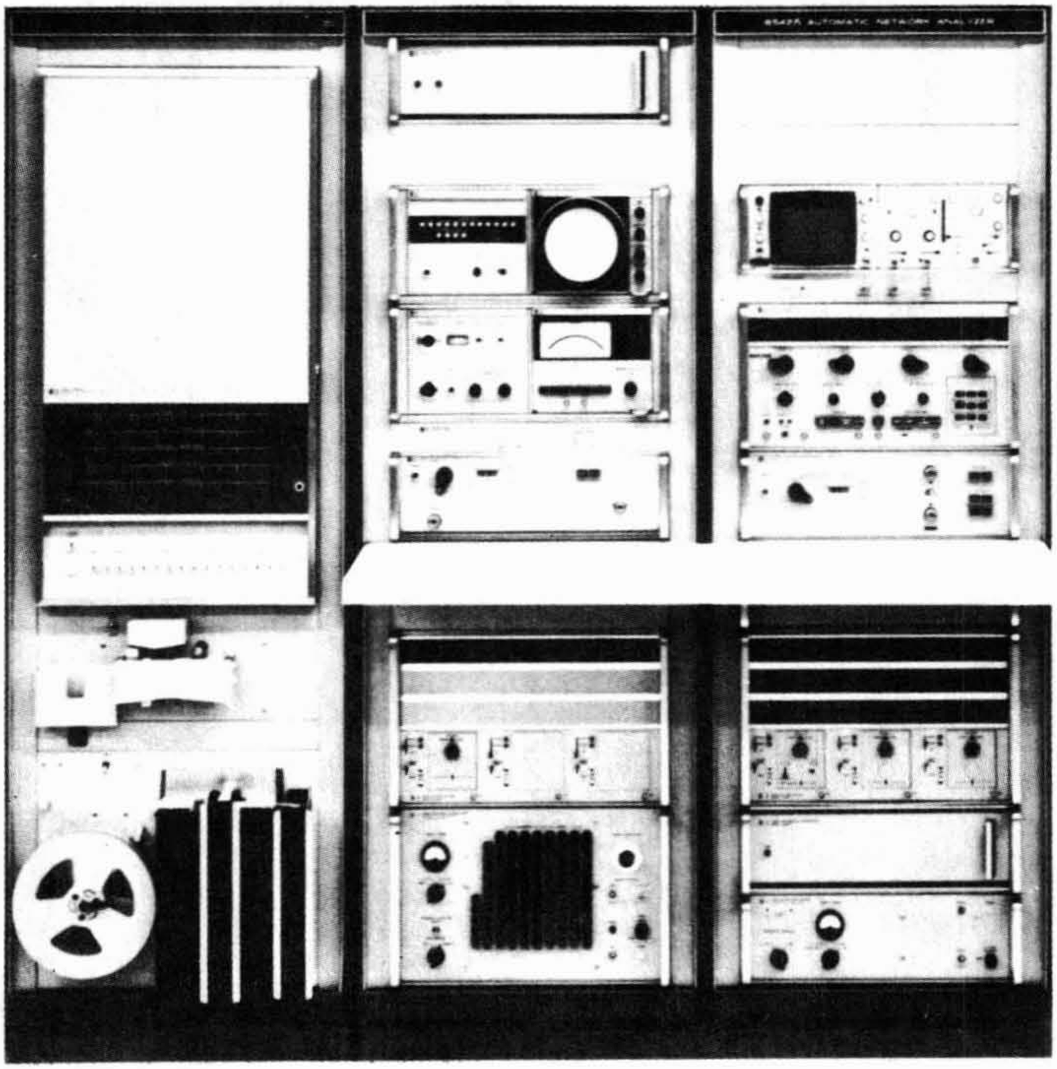
η_s = **power-transfer efficiency of a singly-loaded single-tuned LC network =** $\left[1 - (Q_L / Q_u)\right]$

η_d = **power-transfer efficiency of a doubly-loaded, single-tuned L-C network.** $\left[1 - (Q_L / Q_u)\right]^2$

SECTION 3



RF CAPACITOR CHARACTERIZATION: ACTUAL TEST DATA



INTRODUCTION

As an aid to the design engineer, ATC has developed a wideband test fixture to measure the (series-element) VSWR, I.L. and S-Parameters of a range of capacitance values likely to be highly useful to the engineer The chips (ATC 100, case A) were mounted on 25 mil microstrip lines on 25 mil alumina, and the data obtained on a Hewlett-Packard 8542A Network Analyzer (computer controlled; synthesizer referenced; phase locked; line-printer output).

The data on the following pages is a photocopy of the computer print-out of the Insertion Loss, VSWR, and S-Parameters for capacitance values spaced from 10. pF through 1000.pF. The data is unique in that actual operating parameters are now available over the frequency range of use.

reprinted from:
MICROWAVE SYSTEMS NEWS
JANUARY/FEBRUARY 1971

MICROWAVE AMPLIFIER DESIGN DATA

Capacitor S-Parameters Versus Frequency in Graphic Form

Vincent Perna
Vice President, Microwave Engineering
American Technical Ceramics
Huntington Station, L.I., N.Y. 11746
516-271-9600

Until the recent characterization of capacitors by ATC in S-Parameters at microwave frequencies, the only data typically available was Dissipation Factor in the general region of 1kHz and 1 MHz, with some extrapolated estimates to 100 MHz. Unfortunately, not many microwave amplifiers are built at these frequencies. Additionally, that data is based on a thin-plate or the slug of dielectric material alone, — not that of actual multilayer capacitive structure. However, if actual capacitors are measured in terms of their S-Parameters as a series element on a transmission line, they may be characterized in considerable detail well up into the gigahertz regions.

This "Scattering Parameter" characterization is especially useful up around 1000 MHz, since h-, y-, or z-Parameters require the measurement of voltages and currents . . . a task increasingly difficult at UHF and above.

Measurement as a series element on a 50Ω line with a Hewlett Packard 8542A Network Analyzer has the additional advantage of providing much greater accuracy than that of a high VSWR, shunt-mode measurement at the end of a transmission line.

For a bilateral, passive device, such as an ATC capacitor with its short path-length, only *two* of the four normally necessary parameters are required: the Transmission Coefficient, S_{21}, and the Reflection Coefficient, S_{11}; (other capacitor physical configurations, such as oblong shapes, would require that all four be measured and provided).

Some representative parameters are displayed in the accompanying set of graphs, which are vector quantities displayed here as a magnitude and an angle. They could easily be displayed as two vector components and their signs to show quadrant.

When used as a DC-block on a transmission line, the transmission coefficient, S_{21}, may be used to determine signal attenuation in dB from the equation shown in figure 1. Note that the Transmission Loss versus frequency graph begins at 300 MHz. Below this, the reflected power due to high capacitive reactance dominates. In the graph of S_{21} (fig. 2), a value of 1.00 signifies 100.% signal transmission. No reflection, dissipation, or radiation losses are indicated.

An S_{21} *angle* (figure 3) of only a few degrees over a wide frequency range signifies a capacitor especially suitable for both radar and short rise-time pulse transmission due to the low pulse-shape distortion that would occur. This is due to the differential phase shift of its various Fourier components versus frequency.

The reflection coefficient, S_{11}, is essentially equivalent to the microwave engineer's "gamma" (Γ), and from it may be determined, (thru charts or calculations), the equivalent amount of power reflected before it ever could enter the capacitor to be transmitted.

The interrelationships are given by:

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|} = \frac{1 + |S_{11}|}{1 - |S_{11}|}$$

The best possible VSWR is none at all, but this would require an S_{11} of zero, which is almost impossible to attain. Capacitor VSWR versus frequency for a range of commonly used values is displayed in figure 4.

This reflected power, effectively subtracts from the original signal amplitude, and is shown in figure 5. It may be readily solved for as shown in the "Microwave Engineer's Handbook", (Horizon House).

S-Parameter measurement techniques can be applied to any circuit interconnection or component. Where a very wideband DC-block is required on a transmission line, it will tell in advance a great deal about the suitability of a particular choice and its exact characteristics over any given frequency range. Where load resistance levels are low, a capacitor used in this application must have a very low equivalent series resistance, otherwise the transmission loss will be high. In fact, a low S_{21} value *implies* a high equivalent series resistance at frequencies where S_{11} is low.

A capacitance value that is too small for a DC-block will produce a high VSWR (high S_{11} magnitude). However, for impedance matching applications this is not a restriction.

A too inductive capacitor will have a large variation of the transmission coefficient versus frequency above the range where the VSWR had finished its normal original steep drop off to a low value.

The data which has been briefly presented here provides the design engineer for the first time with significant capacitor characteristics, graphically displayed for ease in visualizing parameter variations and interrelationships versus frequency for specific capacitance values. For more complete information, including full-scale graphs, contact American Technical Ceramics.

Transmission Loss (dB) = Insertion Loss - VSWR Loss

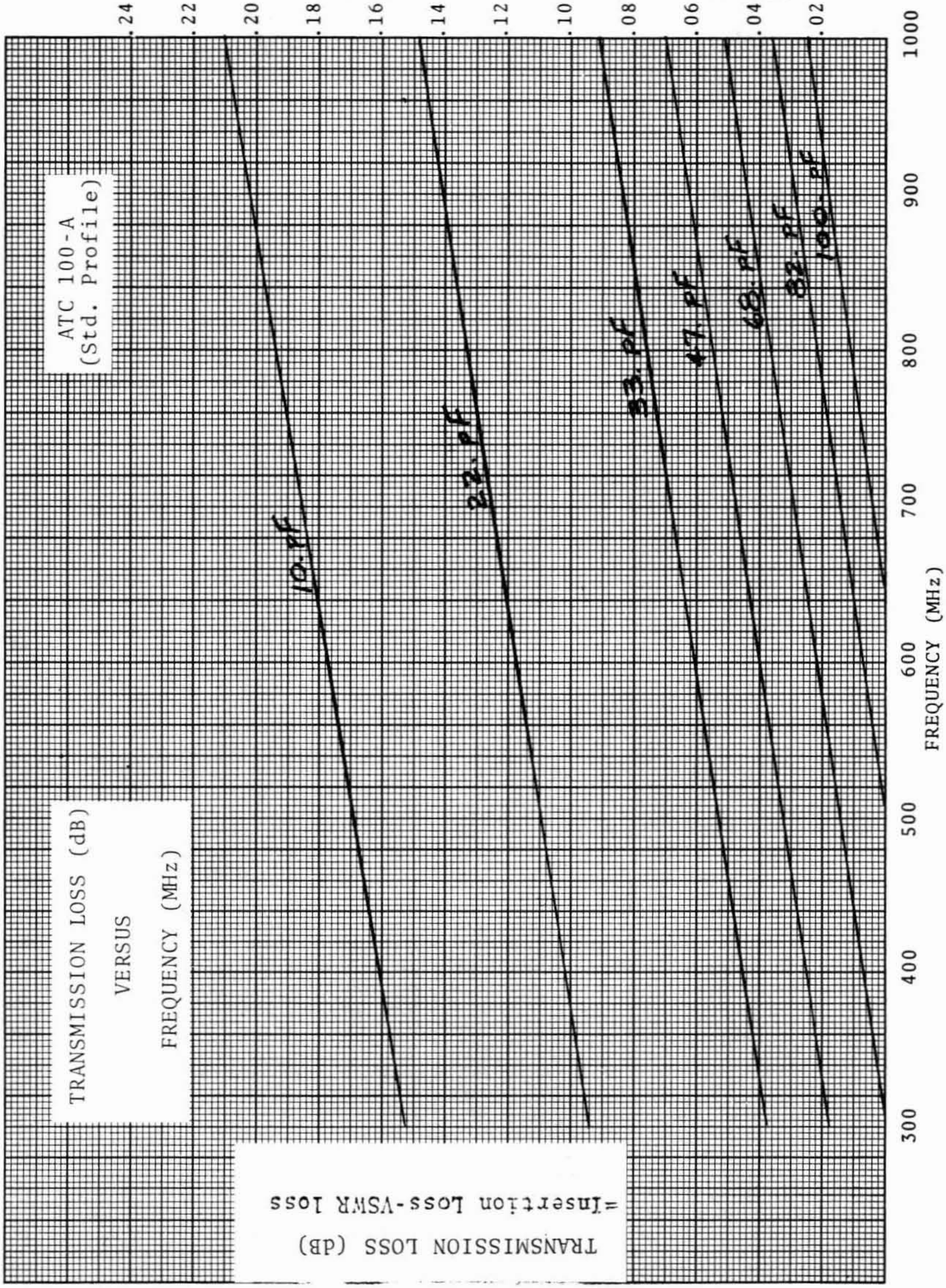


Fig. 1

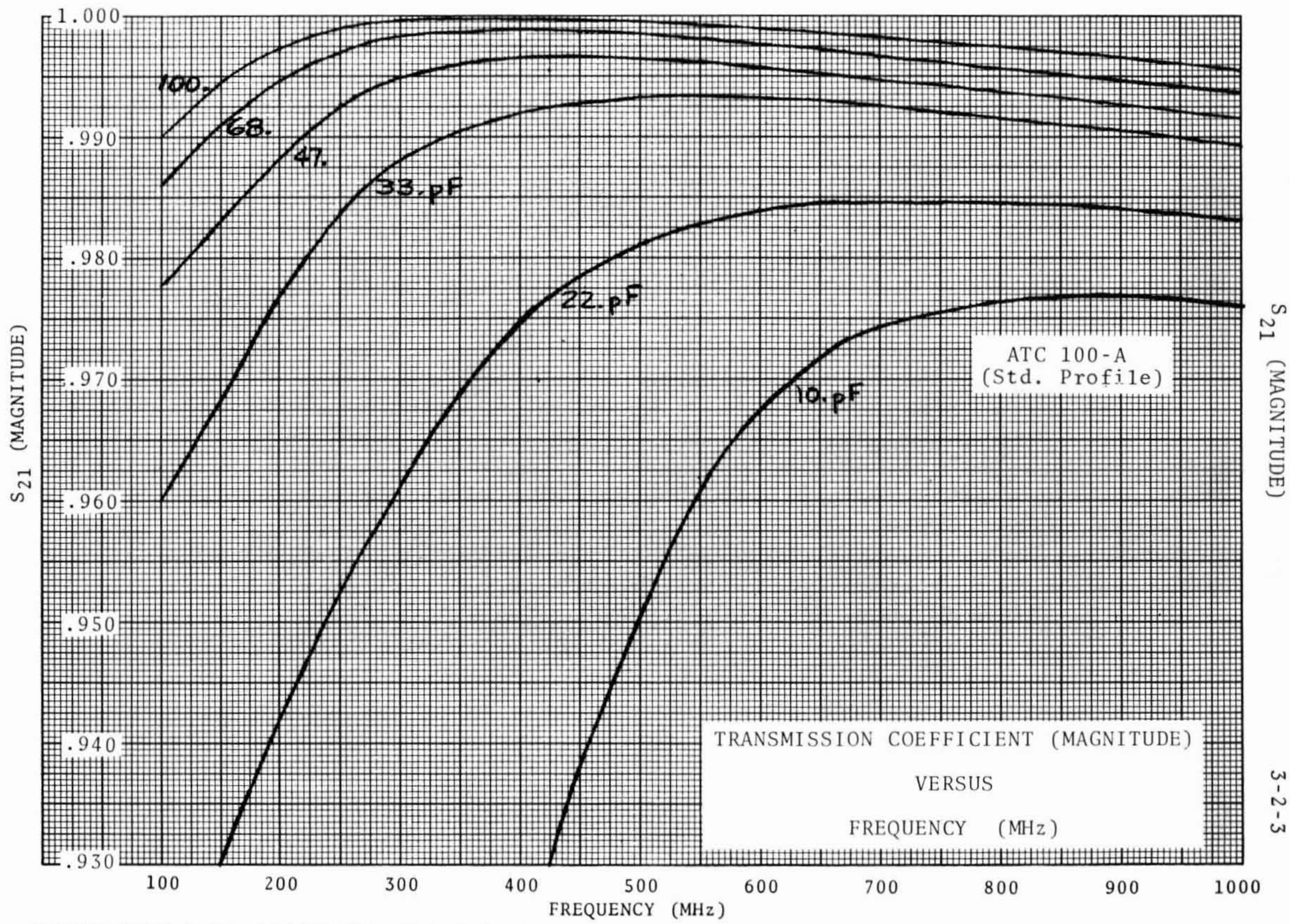


Fig. 2

MICROWAVE AMPLIFIER DESIGN DATA

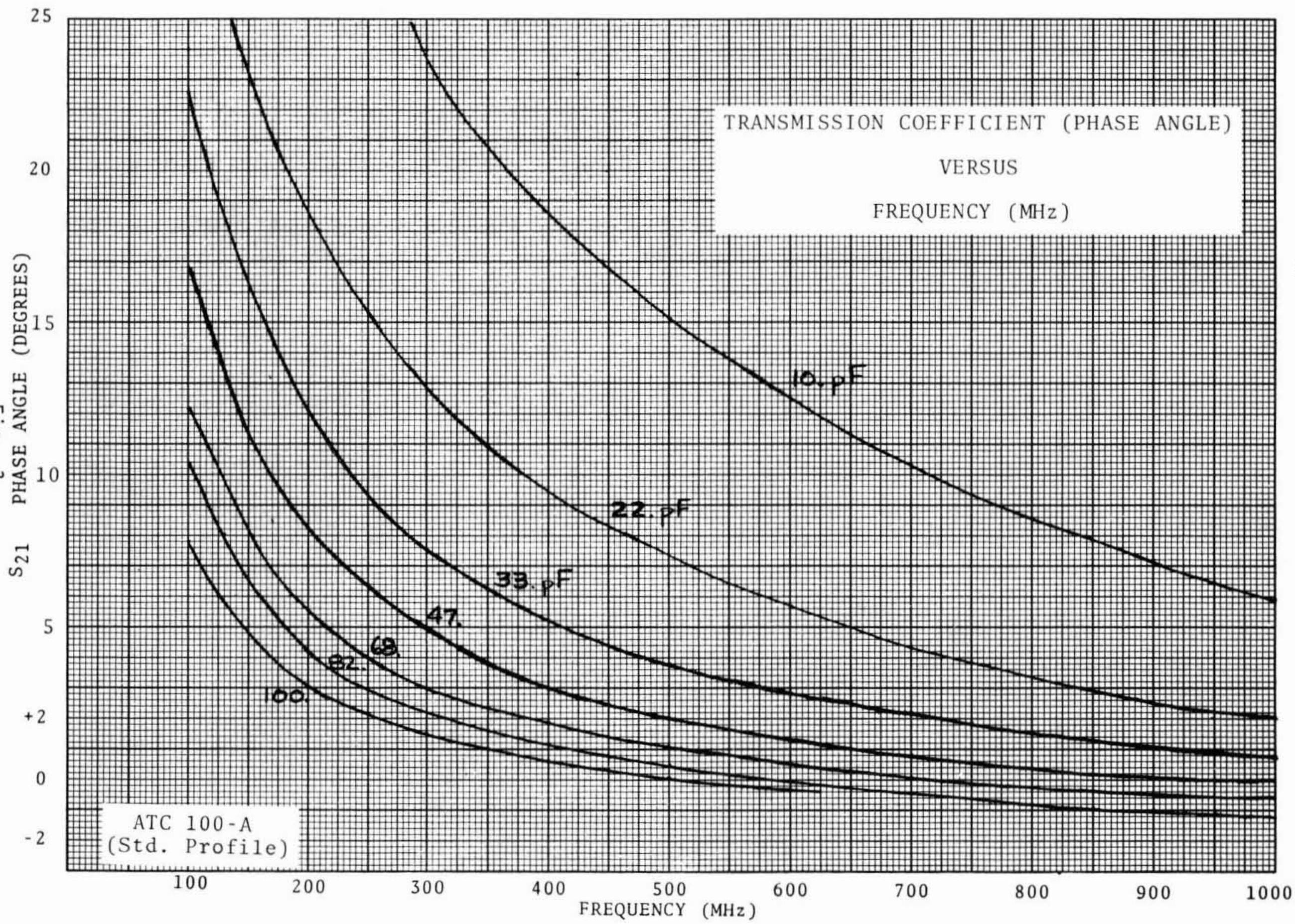

TRANSMISSION COEFFICIENT (PHASE ANGLE)
VERSUS
FREQUENCY (MHz)
S_{21} PHASE ANGLE (DEGREES)
FREQUENCY (MHz)
25
20
15
10
5
+2
0
-2
100
200
300
400
500
600
700
800
900
1000
10. pF
22. pF
33. pF
47.
68.
82.
100.
ATC 100-A
(Std. Profile)

Fig. 3

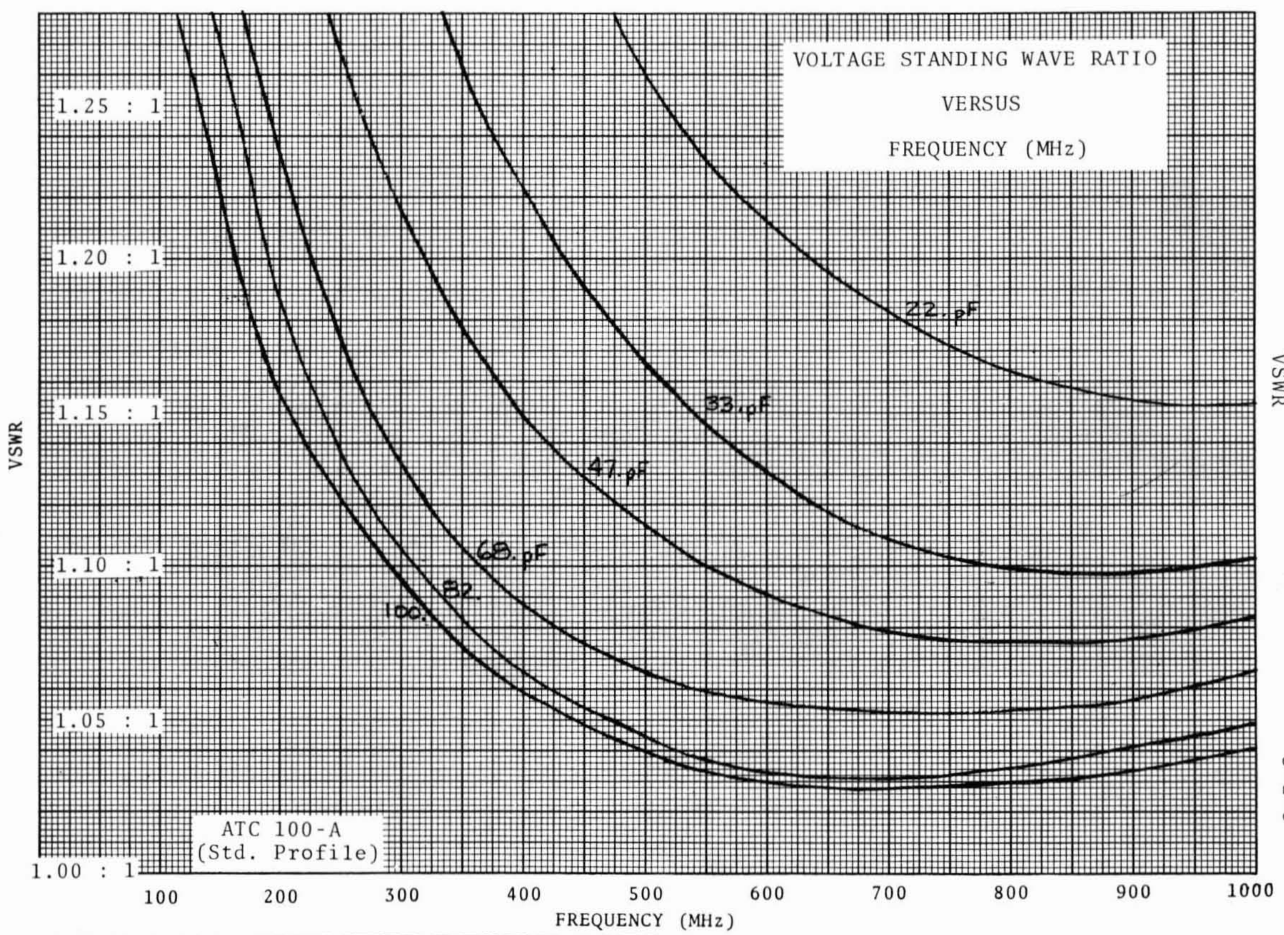
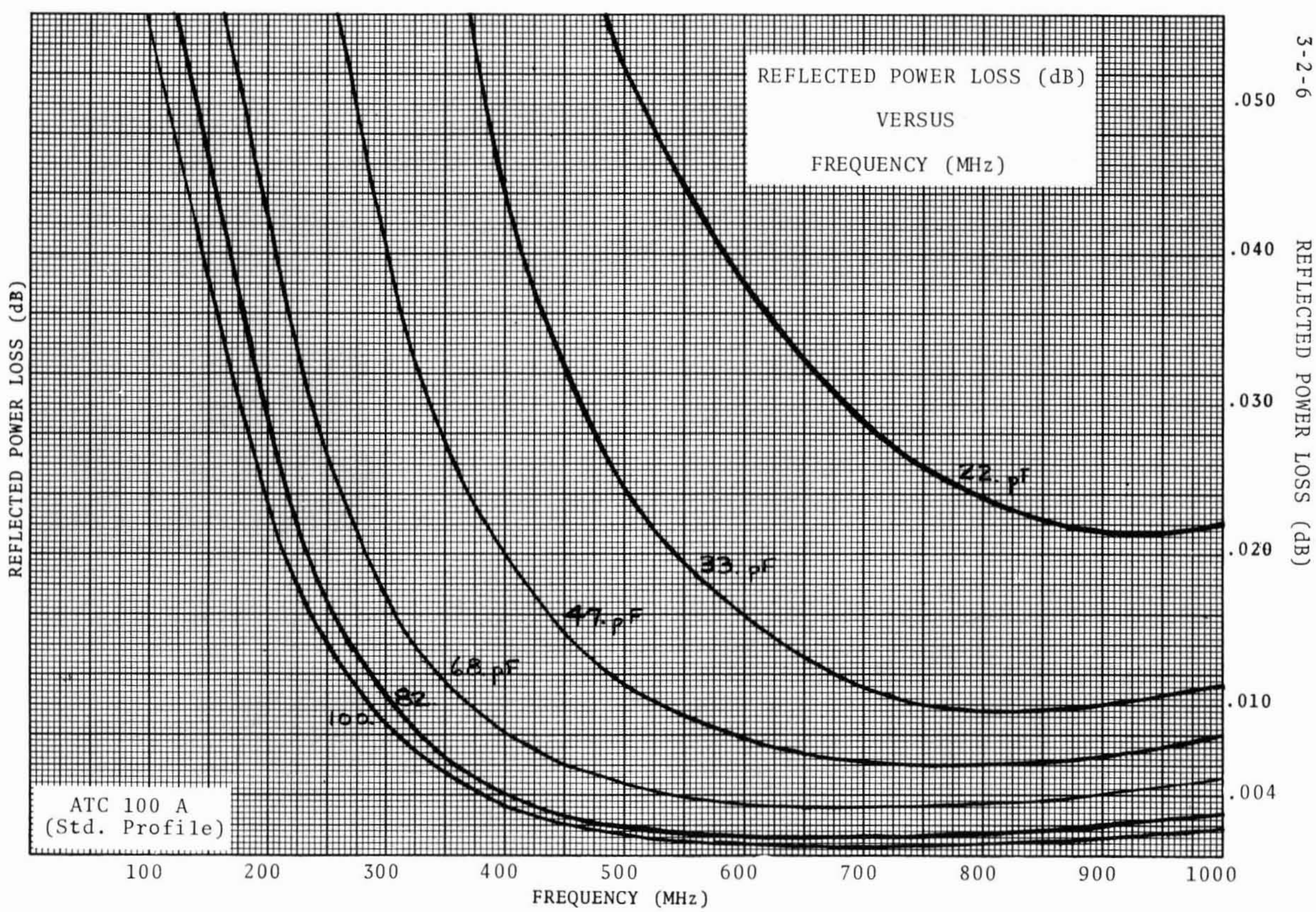


Fig. 4

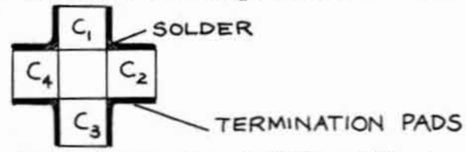
MICROWAVE AMPLIFIER DESIGN DATA


REFLECTED POWER LOSS (dB)
VERSUS
FREQUENCY (MHz)
REFLECTED POWER LOSS (dB)
.050
.040
.030
.020
.010
.004
22 pF
33 pF
47 pF
68 pF
82
100
ATC 100 A
(Std. Profile)
100
200
300
400
500
600
700
800
900
1000
FREQUENCY (MHz)

TESTING FOR RESONANCE CHARACTERISTICS

For many years, when the "self-resonance" of capacitors was spoken of, what was often meant was the series resonance of the capacitor with its leads. A convenient, rough lab-verification of the series-resonant characteristics of leaded capacitors was generally performed with a grid-dip meter or with a reactance bridge.

Reproducing this original method in recent tests at ATC we noted that the highest capacitance unit (1000 pF) in the style with the highest lead inductance (axial wire) resonated at over 200 MHz with approximately 1/8 inch long leads shorted by a bar. Attempts to shorten the leads more than this made it impossible to couple sufficient energy from the grid dip meter to obtain a reliable resonance indication. A greatly improved way to test leadless units with a grid dip oscillator, and one which gives excellent results, is to connect four equal, or nearly so, chips in series so as to form a loop. This is done by soldering the capacitors at right angles at the terminations. Thus this "loop" in the form of a cross gives excellent grid dip coupling while minimizing the mutual inductance between the capacitors. See Fig. 1.



(FIG. 1)

The advantage of this system is that there is no "direct" connection to the capacitor under test, provided minimum grid dip coupling is used. The series resonant frequency of a leadless ceramic capacitor can be quite precisely established if a frequency counter is connected to the grid dip oscillator. Similarly, since the low frequency series capacity of the loop (1/4 of a single unit) and the resonant frequency are known, the series inductance of each capacitor can easily be calculated.

As the RF power and frequency capabilities of transistors have risen well into the Gigahertz region, circuit techniques have advanced to the point where minimum size is now necessary and leads must be dispensed with. Furthermore, at these frequencies test fixtures become critical.

To study the inner workings of chip capacitors at microwave frequencies, ATC had to develop new methods of testing. These tests are comprised basically of an examination of the component as a DC-block on a transmission line for its Insertion Loss, VSWR, and S-Parameter Characteristics. They may be performed on a Hewlett-Packard 8542A Network Analyzer with the capacitors mounted across a 20 mil gap in a 25 mil wide line on alumina. Extraneous reactances introduced by inadequate test fixtures are thereby reduced.

Testing in this manner and observing the Insertion Loss of the capacitor, allows an indication of the self-resonant frequency of the basic component with no effects ascribable to leads. These tests disclosed that, with the capacitor mounted on the top of the substrate with its internal electrodes parallel with the substrate, a series of harmonically-related frequencies were found at which there was an absorptive type of resonance. These points are typically above the frequency of series resonance of the unleaded chip, and all act as if they were parallel-resonant in nature, since the VSWR rises at that point (as well as the loss increasing, due apparently to internal Q-multiplication of current). The effect is relatively narrow-band, since the capacitors have very high Q.

If the capacitor is physically rotated 90° about the axis of the transmission line (so that the internal electrodes are now vertical with respect to the surface of the transmission line), one half of these "resonances" disappear: the "fundamental" and all odd-order harmonics (3rd, 5th, etc.). We attribute this to a reduction in the phase differential of currents passing thru various sections of the capacitor (since now the RF current reaches all of the plates simultaneously).

Where the capacitor size (as in our Case A units) or lead-style precludes easy identification of the correct orientation, a green marking is offered by ATC for the surface of the capacitor which should be located parallel to the surface of the substrate (to indicate "this end up"). This also allows proper orientation with respect to the current in other circuit mountings. Typical readings of parallel-resonant frequency obtained by this method are included in Section 3-3(b).

Some test-fixturing influences remain. The width of the gap in the transmission line can significantly influence the indicated resonant frequency. The exact relationship is not clear yet, since different customers have reported different directions of frequency shift. For example: in one case, in the low UHF region, reducing the gap from 20 mils down to 5 mils shifted the average resonant frequency upwards by about 10%. Much higher frequency tests by another customer had opposite results. The reasons for the difference have not yet been fully determined but several possible influences may be at work: (a) the commonly accepted model of the gap as a shunt-series-shunt set of capacitances in combination with the path length thru the capacitor, (b) the influence on wave propagation of the physical discontinuity introduced by the gap-capacitor pair, (c) the apparent electrical length (reactance versus frequency) of the two line "stubs" under the capacitor in shunt with the ends of the capacitor at higher frequencies, etc.

It is anticipated that further tests will clarify the data. When definitive results become available, ATC will publish them.

Although we have found the S-Parameter tests quite valuable, we also realize they require special costly lab equipment. For establishing the series resonant frequency of leadless ceramic capacitors, the above explained grid dip oscillator system is accurate and convenient for frequencies up to 1 GHz.

Resonance-Free Frequency Ranges

"Parallel"-Resonant Frequencies (MHz) (approx. potential variation: $\pm$7.0%)
(as DC block on 50 Ω line)

HORIZONTAL	VERTICAL
ATC 100-B-102-J-C	
370. MHz	*
625.	625. MHz
900.	*
1150.	?
1500.	*
1750.	?
ATC 100-B-471-J	
525. MHz	*
900.	925. MHz
1300.	*
1650.	1700.
2050.	*
2400.	2550.

Note: Lower capacitance values have resonances above 3.0 GHz.
* = resonance-free in between.

HORIZONTAL	VERTICAL
ATC 100-B-201-J	
825. MHz	*
1400.	1400. MHz
2000.	*
2550.	2550.
ATC 100-A-101-J	
1700. MHz	*
2700.	2800. MHz
(4100.)	*
(5700.)	(5700.)
(7000.)	
ATC 100-A-680-J	
2000. MHz	*
(3000+)	(3000+)
ATC 100-A-470-J	
2750. MHz	*

SERIES-ELEMENT I.L., VSWR, AND S-PARAMETERS

1741
19 SEPT 70

AMERICAN TECHNICAL CERAMICS TEST REPORT
ATC100-A-100-J-C-50 MICROWAVE PORCELAIN
(GREEN) DOT

FREQ MHZ	VSWR	LOSS DB	(RE) S11	(IM) S11	(RE) S21	(IM) S21
100,00	11,86	5,50	,54	-,65	,29	,44
110,00	10,12	4,91	,47	-,67	,33	,46
120,00	9,16	4,39	,41	-,69	,37	,47
130,00	8,18	3,95	,36	-,70	,41	,49
140,00	7,06	3,60	,30	-,69	,44	,50
150,00	6,30	3,32	,25	-,68	,47	,50
160,00	5,72	3,09	,20	-,67	,50	,49
170,00	5,23	2,86	,15	-,66	,53	,49
180,00	4,87	2,65	,11	-,65	,55	,49
190,00	4,58	2,45	,07	-,64	,58	,48
200,00	4,30	2,23	,04	-,62	,61	,48
210,00	4,02	2,11	,01	-,60	,63	,47
220,00	3,78	1,98	-,02	-,58	,65	,46
230,00	3,60	1,87	-,05	-,56	,67	,45
240,00	3,46	1,77	-,08	-,55	,68	,45
250,00	3,33	1,66	-,10	-,53	,70	,44
260,00	3,21	1,48	-,12	-,51	,72	,44
270,00	3,09	1,40	-,14	-,49	,73	,43
280,00	2,97	1,29	-,16	-,47	,75	,42
290,00	2,89	1,23	-,17	-,45	,76	,42
300,00	2,78	1,20	-,19	-,43	,77	,41
310,00	2,71	1,10	-,20	-,41	,78	,40
320,00	2,65	1,02	-,21	-,40	,80	,39
330,00	2,57	,94	-,22	-,38	,81	,39
340,00	2,51	,89	-,23	-,36	,82	,38
350,00	2,44	,85	-,24	-,34	,83	,37
360,00	2,39	,81	-,25	-,32	,84	,36
370,00	2,35	,78	-,26	-,31	,85	,35
380,00	2,30	,74	-,27	-,29	,85	,34
390,00	2,26	,70	-,27	-,27	,86	,33
400,00	2,22	,66	-,28	-,26	,87	,33
410,00	2,17	,67	-,28	-,24	,87	,32
420,00	2,14	,67	-,28	-,23	,87	,32
430,00	2,10	,66	-,29	-,21	,87	,31
440,00	2,07	,63	-,29	-,19	,88	,30
450,00	2,04	,59	-,29	-,18	,88	,30
460,00	2,01	,57	-,29	-,17	,89	,29
470,00	1,99	,55	-,29	-,15	,89	,29
480,00	1,97	,54	-,29	-,14	,90	,28
490,00	1,94	,53	-,29	-,13	,90	,28
500,00	1,92	,53	-,29	-,11	,90	,28
525,00	1,87	,48	-,29	-,09	,91	,26
550,00	1,82	,50	-,28	-,06	,91	,25
575,00	1,78	,45	-,28	-,03	,92	,25
600,00	1,74	,45	-,27	-,01	,92	,24
625,00	1,71	,38	-,26	,01	,93	,23
650,00	1,68	,36	-,25	,03	,93	,22
675,00	1,65	,35	-,24	,05	,94	,21
700,00	1,62	,33	-,23	,07	,94	,20
725,00	1,61	,30	-,22	,08	,95	,20
750,00	1,59	,28	-,20	,10	,95	,19
775,00	1,56	,27	-,19	,11	,95	,18

Complete data as illustrated on this page is available by contacting American Technical Ceramics.

SHUNT-ELEMENT REACTANCE, ADMITTANCE, AND REFLECTION COEFFICIENT

A Smith Chart plot of a capacitor's reactance versus frequency will also show its increasing equivalent series resistance (by movement in toward the center of the chart), its series resonant frequency (by crossing the horizontal centerline from the bottom half into the top half of the chart), and points of parallel resonance (by "loops").

Perhaps the most interesting feature is the reversal of direction of the trace exhibited in the loops. The reason for it may be seen in Figure 1 (below), where the change if direction displayed by the trace is quite apparent.

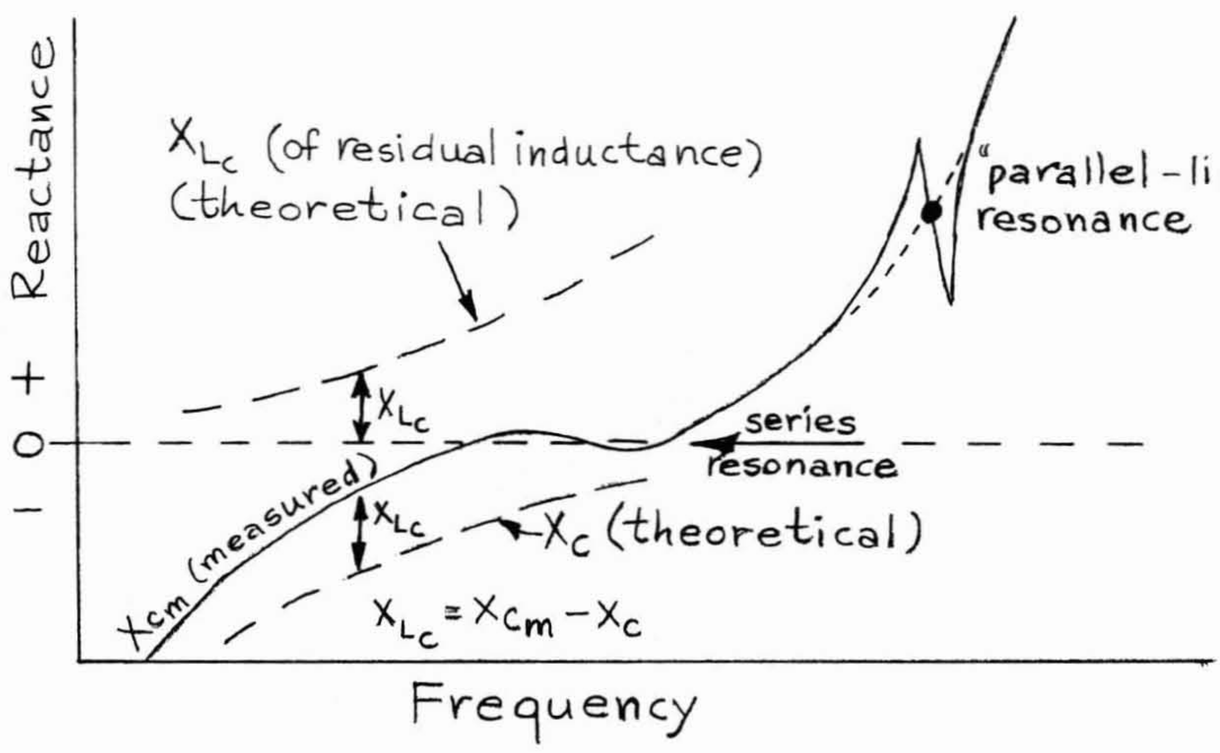


Figure 1

The series inductance, L_s, progressively cancels some of the capacitive reactance, so that, when approaching series resonance, a 43.0 pF capacitor could (reactively) act more like a 1000. pF unit, that is, the effective capacity, C_e, varies by factor $K = \frac{1}{1-\omega^2 L_s C_o}$ where C_o is the low frequency capacity. This is one reason for some of the difficulty experienced at UHF when trying to convert a paper-design to actual hardware, especially in broadband amplifiers.

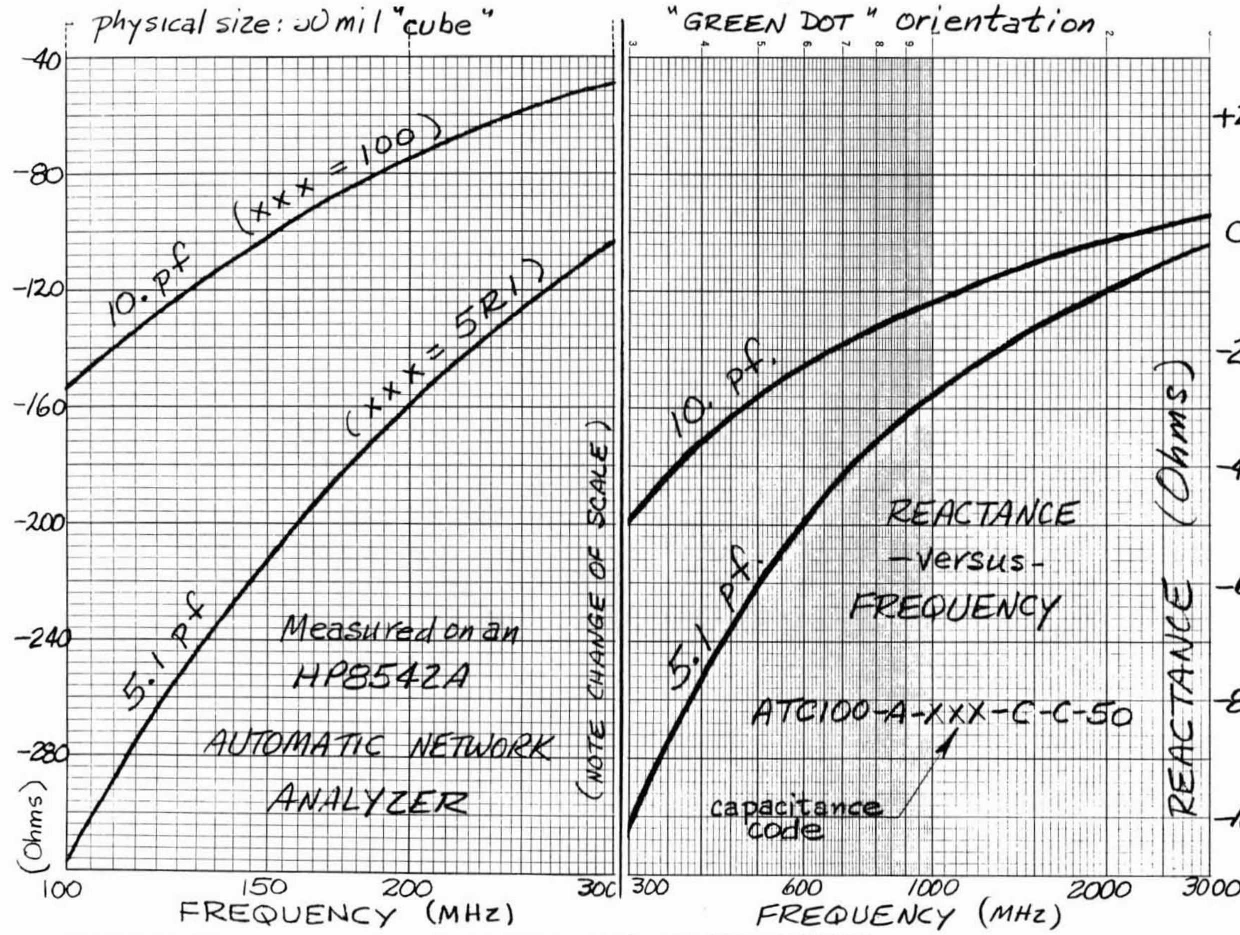
To improve this situation, charts of measured reactance-versus-frequency-versus capacitance value-versus-physical size were made (and are included here). With them, the engineer may select the capacitor which provides the required effective reactance he needs, in the physical space available, and specify the Part Number (for his Bill of Materials) of whatever capacitor it is that provides it. (Typically, the nominal value of the capacitor will be less than what the pure mathematics of the circuit design would dictate.)

The reactance curves have been slightly changed to correct for amplitude ripples inherent in the network analyzer response in order to demonstrate the general trend of measured reactances.

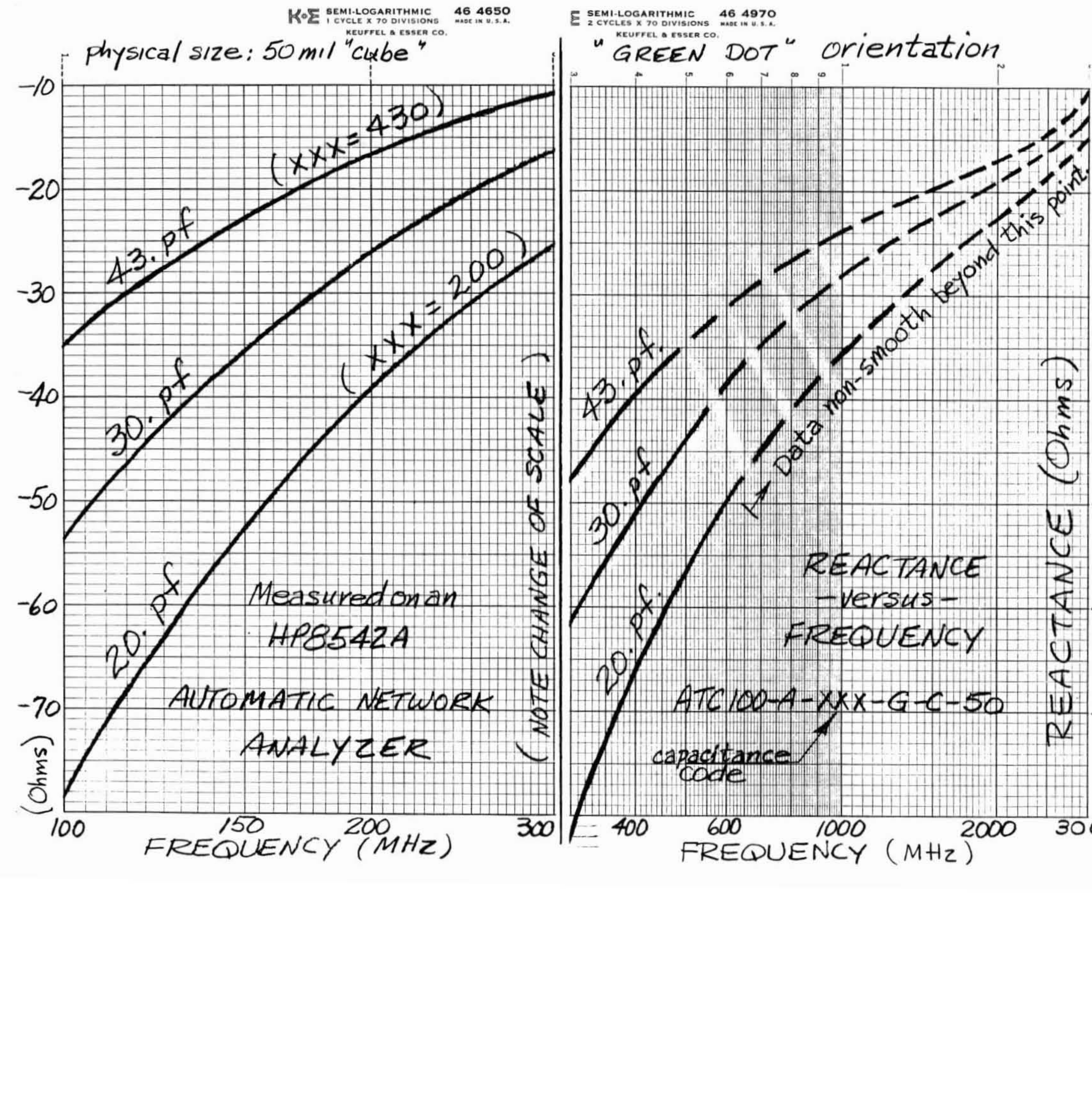
To signify the approximate center frequency of the parallel-like resonance, a black dot is placed on the reactance plot.

As annotated on the graphs, this data was obtained from measurements in which the capacitor was physically oriented with respect to the current flow in such a manner that all of the current reached all of the internal electrodes essentially simultaneously (vertical orientation).

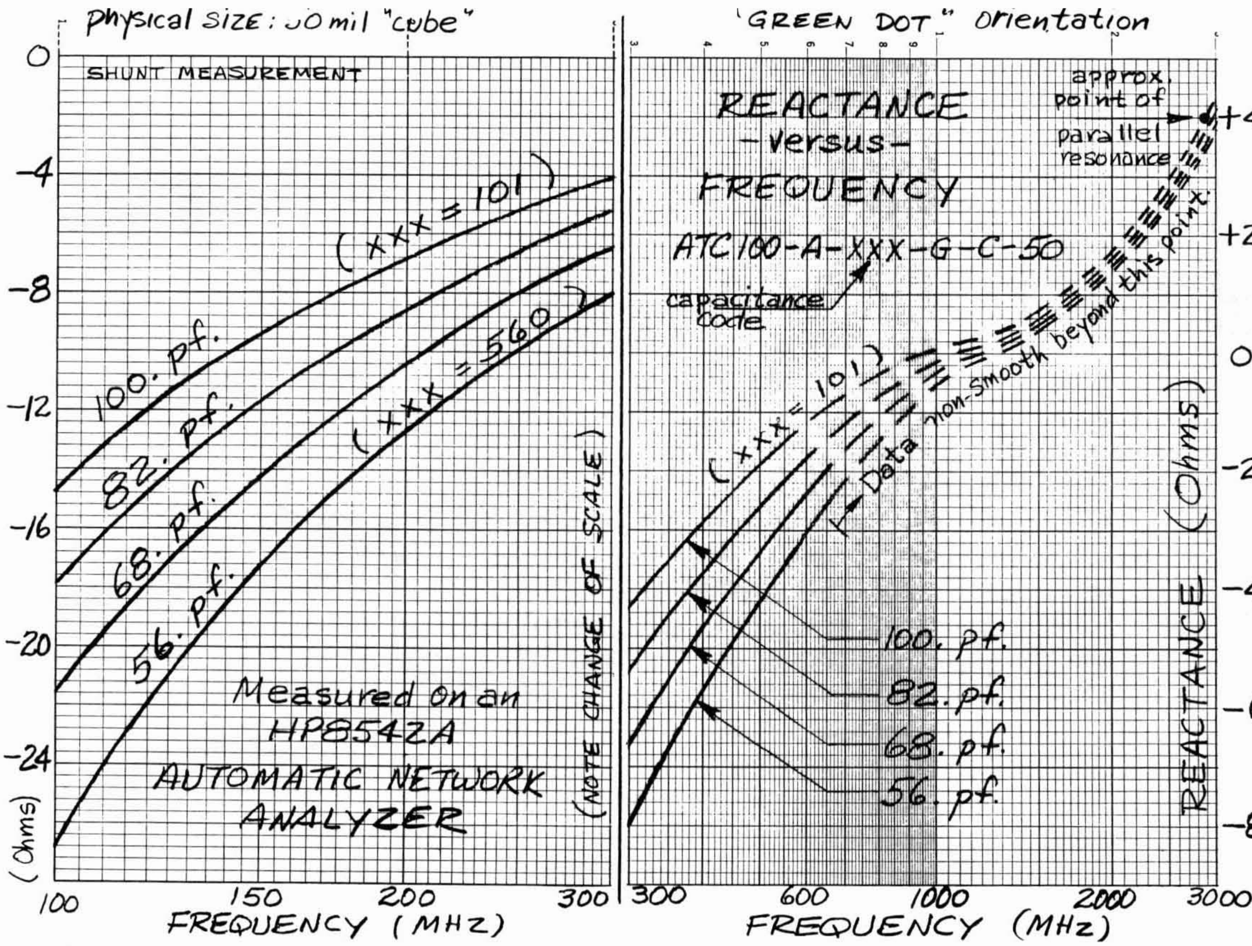
In addition to the smoothed, graphical presentation of the reactance, the raw data is also provided so that a preview of other useful design information is available to the engineer.



SHUNT-ELEMENT REACTANCE-VS-FREQUENCY

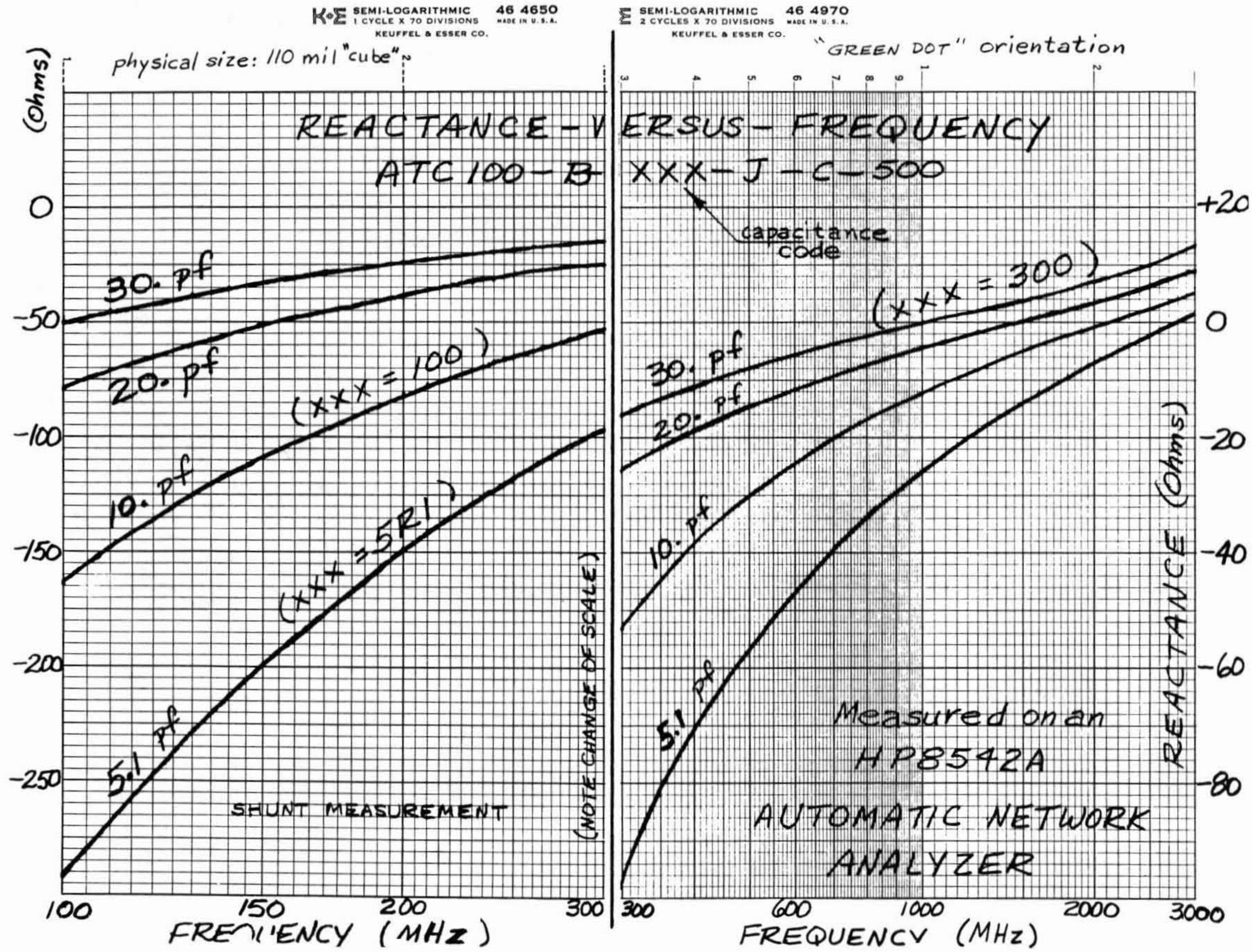

K+E SEMI-LOGARITHMIC 1 CYCLE X 70 DIVISIONS 46 4650 MADE IN U.S.A. KEUFFEL & ESSER CO.
E SEMI-LOGARITHMIC 2 CYCLES X 70 DIVISIONS 46 4970 MADE IN U.S.A. KEUFFEL & ESSER CO.
physical size: 50 mil "cube"
"GREEN DOT" orientation
43. pf (XXX=430)
30. pf
20. pf (XXX=200)
Measured on an HP8542A
AUTOMATIC NETWORK ANALYZER
(NOTE CHANGE OF SCALE)
Data non-smooth beyond this point.
REACTANCE -versus- FREQUENCY
ATC100-A-XXX-G-C-50
capacitance code
(Ohms)
REACTANCE (Ohms)
-10
-20
-30
-40
-50
-60
-70
+4
0
-4
-8
-12
-16
-20
100
150
200
300
400
600
1000
2000
3000
FREQUENCY (MHz)
FREQUENCY (MHz)

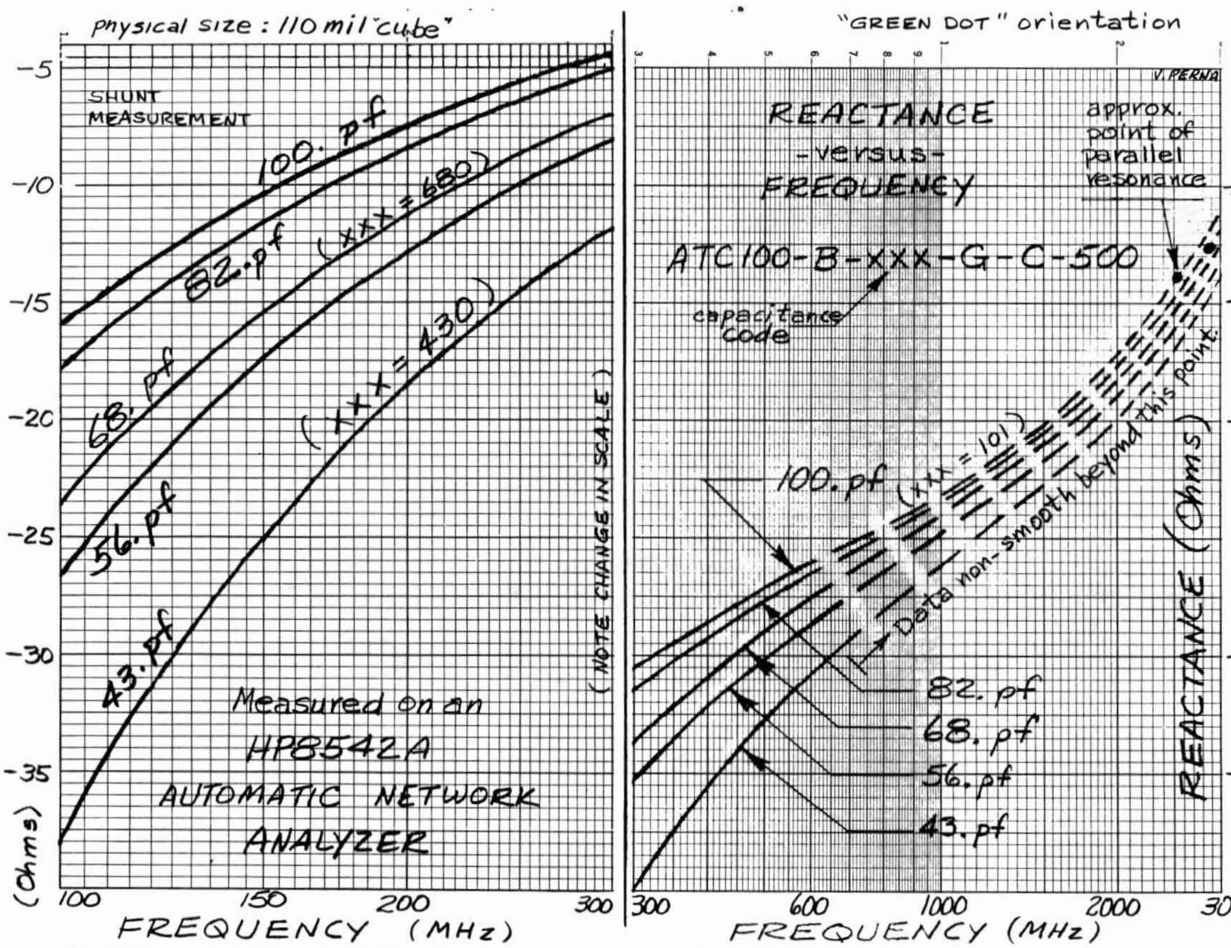
3-5-5



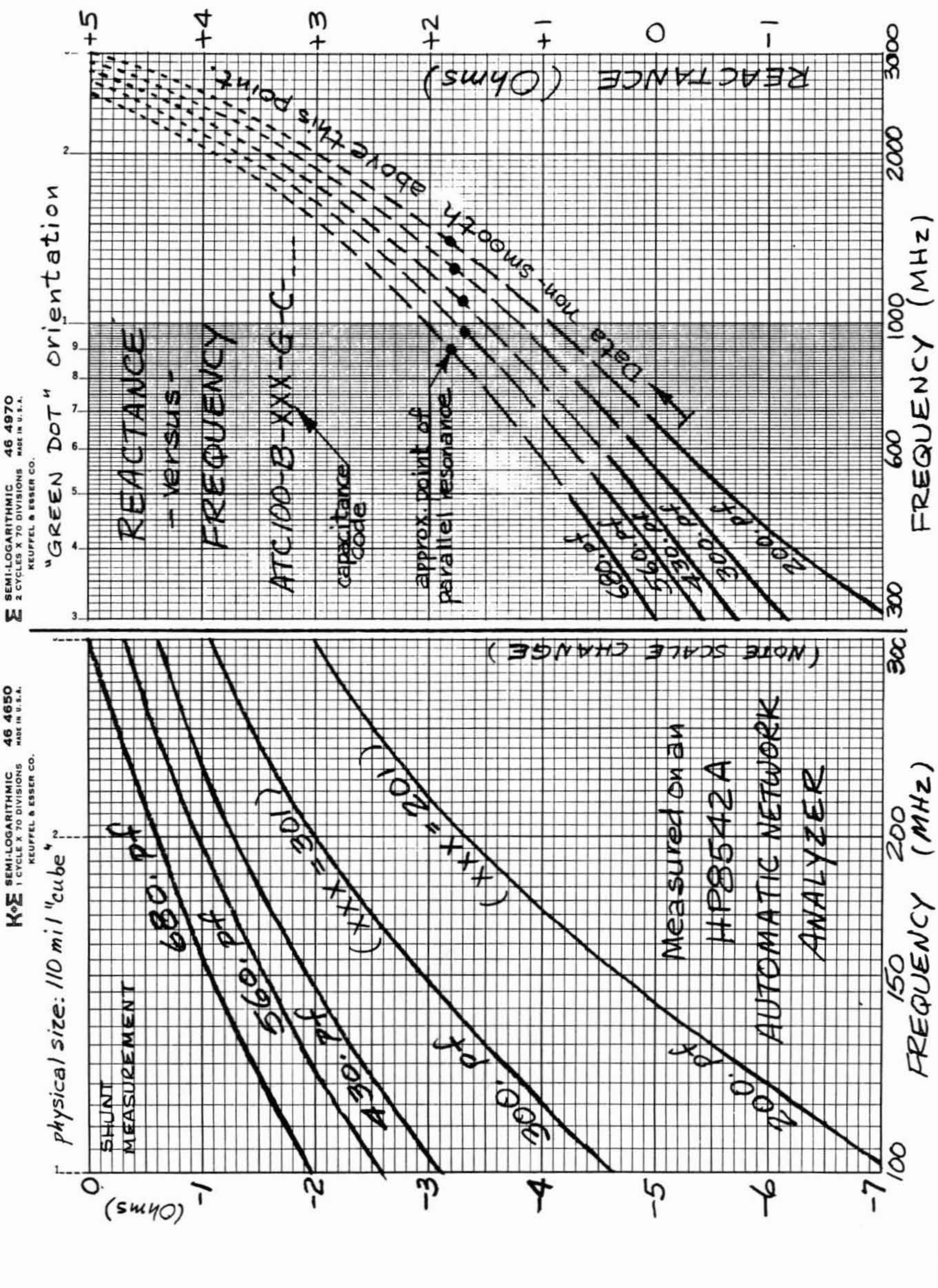
SHUNT-ELEMENT REACTANCE-VS-FREQUENCY

3-5-6

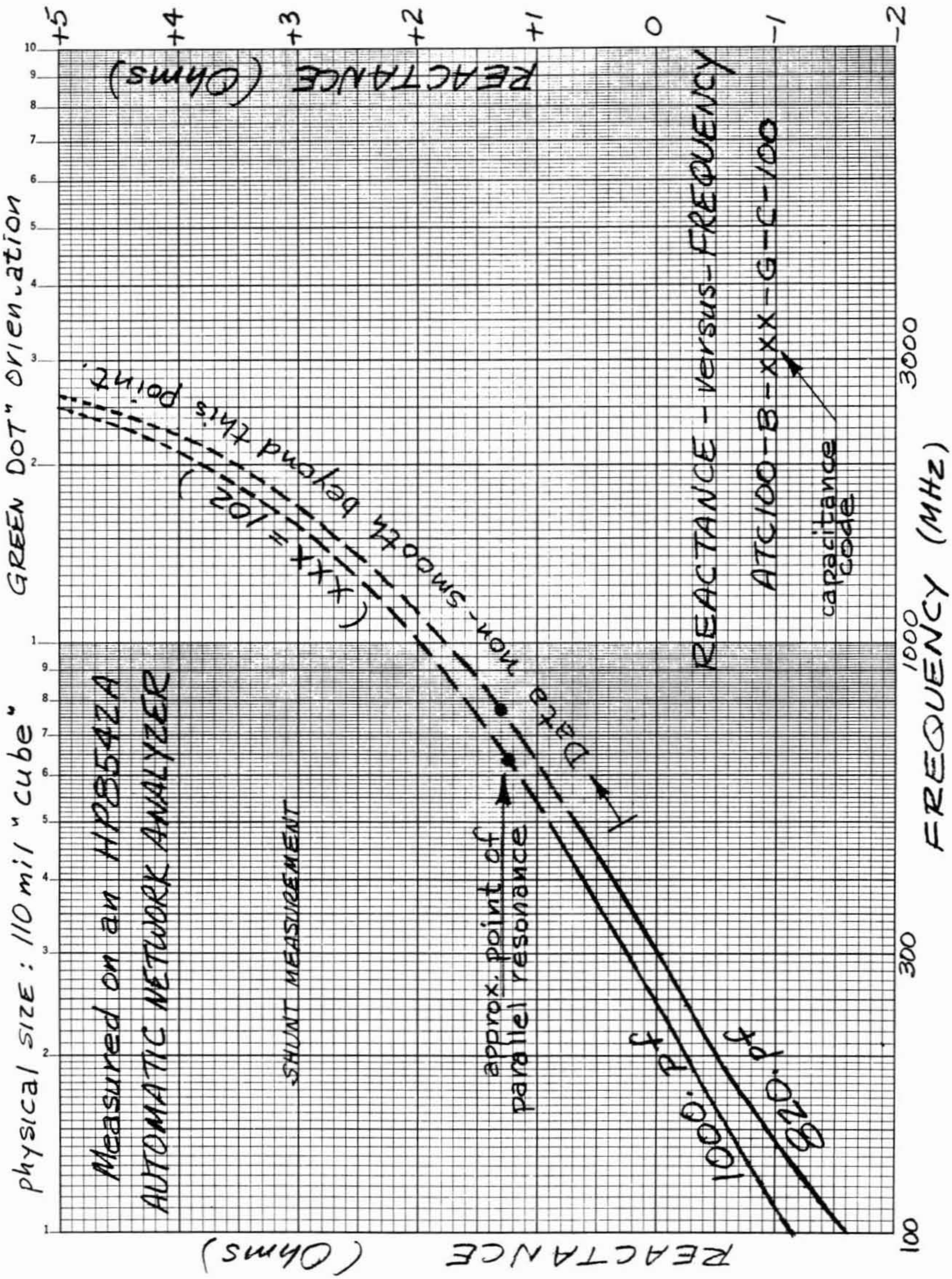

K+E SEMI-LOGARITHMIC 1 CYCLE X 70 DIVISIONS KEUFFEL & ESSER CO. 46 4650 MADE IN U.S.A.
K+E SEMI-LOGARITHMIC 2 CYCLES X 70 DIVISIONS KEUFFEL & ESSER CO. 46 4970 MADE IN U.S.A.
physical size: 110 mil "cube"
"GREEN DOT" orientation
REACTANCE-VERSUS-FREQUENCY
ATC100-B
XXX-J-C-500
capacitance code
(Ohms)
0
-50
-100
-150
-200
-250
30. pf
20. pf
(xxx = 100)
10. pf
(xxx = 5R1)
5.1 pf
SHUNT MEASUREMENT
(NOTE CHANGE OF SCALE)
100
150
200
300
FREQUENCY (MHz)
(xxx = 300)
30. pf
20. pf
10. pf
5.1 pf
Measured on an
HP8542A
AUTOMATIC NETWORK
ANALYZER
REACTANCE (Ohms)
+20
0
-20
-40
-60
-80
300
600
1000
2000
3000
FREQUENCY (MHz)



SHUNT-ELEMENT REACTANCE-VS-FREQUENCY



K+E SEMI-LOGARITHMIC 1 CYCLE X 70 DIVISIONS KEUFFEL & ESSER CO. 46 4650 MADE IN U.S.A.
physical size: 110 mil "cube"
SHUNT MEASUREMENT
680. pf
560. pf
430. pf
300. pf
200. pf
(XXX = 301)
(XXX = 201)
Measured on an
HP8542A
AUTOMATIC NETWORK
ANALYZER
(NOTE SCALE CHANGE)
100
150
200
300
FREQUENCY (MHz)
0
-1
-2
-3
-4
-5
-6
-7
(Ohms)
SEMI-LOGARITHMIC 2 CYCLES X 70 DIVISIONS KEUFFEL & ESSER CO. 46 4970 MADE IN U.S.A.
"GREEN DOT" orientation
REACTANCE
- versus -
FREQUENCY
ATC100-B-XXX-G-C
capacitance code
approx. point of parallel resonance
Data non-smooth above this point
680. pf
560. pf
430. pf
300. pf
200. pf
300
600
1000
2000
3000
FREQUENCY (MHz)
REACTANCE (Ohms)
+5
+4
+3
+2
+1
0
-1



SHUNT-ELEMENT REACTANCE-VS-FREQUENCY

CM 2997
10 APR 72

ATC REPORT, SIMPLE SHUNT FIXTURE GREEN
DOT=PLATES VERT., ATC100-A-XXX-J-C
5.1PF 1

FREQ MHZ	X/ZO	(RE) S11	(IM) S11	X-OHMS	Y-MAG	Y-ANG	G/YO	B/YO	FREQ MHZ
100.000	-6.414	.953	-.305	-320.703	.156	90.121	.000	.156	100.000
110.000	-5.773	.944	-.337	-288.672	.173	90.472	-.001	.173	110.000
120.000	-5.297	.937	-.367	-264.844	.189	90.896	-.003	.189	120.000
130.000	-4.898	.930	-.396	-244.922	.204	91.529	-.005	.204	130.000
140.000	-4.555	.915	-.422	-227.734	.219	91.031	-.004	.219	140.000
150.000	-4.266	.902	-.448	-213.281	.234	90.941	-.004	.234	150.000
200.000	-3.180	.825	-.576	-158.984	.314	90.634	-.003	.314	200.000
210.000	-3.035	.808	-.598	-151.758	.330	90.443	-.003	.330	210.000
220.000	-2.906	.792	-.618	-145.313	.344	90.425	-.003	.344	220.000
230.000	-2.758	.770	-.643	-137.891	.363	90.227	-.001	.363	230.000
240.000	-2.633	.745	-.662	-131.641	.380	89.712	.002	.380	240.000
250.000	-2.496	.721	-.688	-124.805	.400	89.669	.002	.400	250.000
300.000	-2.074	.617	-.775	-103.711	.482	89.340	.006	.482	300.000
325.000	-1.893	.560	-.820	-94.629	.528	89.498	.005	.528	325.000
350.000	-1.766	.508	-.847	-88.281	.566	89.145	.008	.566	350.000
375.000	-1.629	.443	-.872	-81.445	.613	88.543	.016	.613	375.000
400.000	-1.520	.389	-.903	-75.977	.658	88.955	.012	.658	400.000
425.000	-1.434	.336	-.910	-71.680	.697	88.150	.022	.696	425.000
450.000	-1.348	.282	-.930	-67.383	.742	88.255	.023	.741	450.000
475.000	-1.281	.237	-.940	-64.063	.780	88.181	.025	.779	475.000
500.000	-1.213	.186	-.948	-60.645	.824	87.996	.029	.823	500.000
525.000	-1.154	.139	-.956	-57.715	.866	88.002	.030	.865	525.000
550.000	-1.102	.094	-.954	-55.078	.907	87.574	.038	.906	550.000
575.000	-1.053	.050	-.959	-52.637	.950	87.702	.038	.949	575.000
600.000	-1.008	.009	-.957	-50.391	.991	87.515	.043	.990	600.000
625.000	-.967	-.031	-.956	-48.340	1.032	87.451	.046	1.031	625.000
650.000	-.925	-.074	-.956	-46.240	1.081	87.592	.045	1.080	650.000
675.000	-.888	-.112	-.953	-44.385	1.126	87.605	.047	1.125	675.000
700.000	-.852	-.152	-.945	-42.578	1.173	87.447	.052	1.172	700.000
725.000	-.819	-.188	-.940	-40.967	1.218	87.550	.052	1.217	725.000
750.000	-.786	-.225	-.930	-39.307	1.271	87.412	.057	1.270	750.000
775.000	-.757	-.258	-.916	-37.842	1.320	87.048	.068	1.318	775.000
800.000	-.729	-.291	-.907	-36.475	1.369	87.051	.070	1.367	800.000
825.000	-.704	-.321	-.896	-35.205	1.418	87.004	.074	1.416	825.000
850.000	-.682	-.346	-.884	-34.082	1.463	86.781	.082	1.461	850.000
875.000	-.660	-.372	-.873	-33.008	1.512	86.724	.086	1.510	875.000
900.000	-.643	-.394	-.863	-32.129	1.555	86.675	.090	1.553	900.000
925.000	-.622	-.417	-.850	-31.104	1.605	86.471	.099	1.602	925.000
950.000	-.602	-.444	-.839	-30.078	1.659	86.600	.098	1.656	950.000
975.000	-.585	-.465	-.827	-29.248	1.708	86.550	.103	1.705	975.000
1000.000	-.566	-.487	-.813	-28.320	1.761	86.360	.112	1.758	1000.000
1100.000	-.507	-.561	-.767	-25.342	1.967	86.335	.126	1.963	1100.000
1200.000	-.452	-.631	-.717	-22.583	2.211	86.480	.136	2.207	1200.000
1300.000	-.403	-.692	-.668	-20.166	2.477	86.801	.138	2.473	1300.000
1400.000	-.359	-.736	-.607	-17.944	2.777	85.815	.203	2.770	1400.000
1500.000	-.318	-.769	-.545	-15.894	3.126	84.100	.321	3.109	1500.000
1600.000	-.291	-.782	-.499	-14.551	3.404	82.027	.472	3.371	1600.000
1700.000	-.268	-.799	-.461	-13.379	3.692	80.792	.591	3.645	1700.000
1800.000	-.238	-.814	-.411	-11.877	4.122	78.466	.824	4.039	1800.000
1900.000	-.223	-.809	-.381	-11.145	4.339	75.213	1.107	4.195	1900.000
2000.000	-.210	-.827	-.364	-10.486	4.624	75.893	1.127	4.484	2000.000
2100.000	-.189	-.785	-.311	-9.473	4.792	65.231	2.008	4.352	2100.000
2200.000	-.183	-.823	-.313	-9.131	5.156	70.226	1.744	4.852	2200.000
2300.000	-.150	-.857	-.264	-7.495	6.256	69.687	2.172	5.867	2300.000
2400.000	-.133	-.853	-.232	-6.653	6.794	64.673	2.906	6.141	2400.000
2500.000	-.115	-.857	-.200	-5.743	7.595	60.719	3.715	6.625	2500.000
2600.000	-.093	-.859	-.161	-4.626	8.727	53.848	5.148	7.047	2600.000
2700.000	-.077	-.860	-.134	-3.857	9.631	47.992	6.445	7.156	2700.000
2800.000	-.069	-.855	-.119	-3.455	9.921	43.245	7.227	6.797	2800.000
2900.000	-.058	-.871	-.102	-2.908	11.395	41.526	8.531	7.555	2900.000
3000.000	-.043	-.880	-.075	-2.130	13.275	34.401	10.953	7.500	3000.000

Complete data as illustrated on this page is available by contacting American Technical Ceramics.

SECTION 5

ADDITIONAL REFERENCE DATA

RF CIRCUIT DESIGN BIBLIOGRAPHY

1. Benjamin, J.A., "Build Broadband RF Power Amplifiers", Electronic Design 2, January 18, 1969, pp. 50-54.

2. Kamnitsis, C., "Broadband Matching of UHF Microstrip Amplifiers", MicroWaves, April, 1969, pp. 54-56.

3. Froehner, W.H., "Quick Amplifier Design with Scattering Parameters", Electronics, October 16, 1967, pp. 100-109.

4. Lee, H.C. and Hodowanec, G., "The Use of Coaxial-Package Transistors in Microstripline Circuits", RCA Application Note #AN-4025.

5. Matthaei, G.L., "Tables of Chebyshev Impedance Transforming Networks of Low-Pass Filter Form", Proc. of IEEE, Vol. 52, August, 1964, pp. 939-963.

6. Grammer, G. "Simplified Design of Impedance-Matching Networks", QST (ARRL), March 1967, pp. 38-42, also, April and May.

7. Cowles, L.G., "Transistor Circuits and Applications", Prentice Hall, 1968, pp. 308-310.

8. Reference Data for Engineers, ITT Corp., 4th Edition, 1960, pp. 120-127, 583-587.

9. Benjamin, J., "How to Measure Large Signal Parameters of RF Power Transistors", EEE, March, 1970, pp. 96-97.

10. Battisti, R.A., and Morey, K.O., "Amplifier and Oscillator Applications of the Gigahertz Family of UHF Transistors 2N4429, 2N4430, 2N4431, 2N4976, pp. 11, 12.

11. Minton, B., "VHF Amplifier Design", RCA Application Note SMA-36.

12. Valley, Jr., G.E., and Wallman, H., "Vacuum Tube Amplifiers", MIT Rad. Lab. Series, 284-289, Vol. 18, pp. 166-191.

13. Tatum, J., "RF Large-Signal Transistor Power Amplifiers, Part III, Matching Networks", EDN, July, 1965, p. 66.

14. Magill, L.M., "Watt-Megahertz Ratings Run Second to Reliability in Foreign RF Power Transistor", Electronics, April 27, 1970, pp. 80-89.

15. Motorola MM4019 Data Sheet #DS5380 of November, 1969.

16. ITT Semiconductors Data Sheet #3TE609.

17. Davis, F., "Matching Network Designs with Computer Solutions", Motorola Semiconductor Products Inc., Application Note "AN-267.

18. "Composite Transistor Touts 2.5 GHz f_t", EDN, April 15, 1970, p. 26.

19. Emery, F.E., and O'Hagan, H., "Optimal Design of Matching Networks for Microwave Transistor Amplifiers", IEEE Trans. on MTT, (Correspondence Section,) December, 1966, pp. 696-698.

20. Udelson, B.J., and McDonnald, D.W., "The Design of Two-Section Transformers Having Variable Length", The Microwave Journal, April, 1966, pp. 68-70.

21. Ruthroff, C.L., "Some Broad-Band Transformers", Proc. of IRE, August, 1959, pp. 1337-1341.

22. Alexander, W., "Tolerance Limits in Matching", Electronic Engineering ((British)), April, 1956, pp. 162-164.

23. Chasek, N.E., "Broad-Band Transistor Amplifiers", NEREM Record, 1964, pp. 54-55.

24. Allen, P., "Multi-Core Transformers Boost Bandwidth", Electronic Design, February 3, 1964, pp. 30-32.

25. Talkin, A.I. and Cuneo, J.V., "Wideband Balun Transformer", The Review of Sceintific Instruments, Vol. 28, No. 10, October, 1957, pp. 808-815.

26. Howell, D.W., "A Transistor Amplifier with 500. Mc Low-Pass Bandwidth", IEEE Trans. on Circuit Theory (Correspondence), Vol. CT-12, No. 4, December, 1965, pp. 591-593.

27. Jones, B.L., "A 225 Mc/s Amplifier Using Silicon Transistors", Fairchild Semiconductor Corp., Application Note APP-22.

28. Hejhall, R., "A 160 MHz 15-Watt Solid-State Power Amplifier", Motorola Semiconductor Products Inc., Application Note #AN-214.

29. Tatom, C., "Whats and Whys About Y-Parameters", Motorola Semiconductor Products Inc., Application Note #AN-158, p. 6.

30. Glenovatch, V., "Design of Distributed Transistor Amplifiers at Microwave Frequencies", The Microwave Journal, January, 1967, pp. 41-47.

31. Johnson, G., Norris, P., Opp, F., "High-Frequency Amplifier Design Using Admittance Parameters", Electro-Technology, December, 1963, pp. 66-72, 140.

32. Fredricks, G., "How to Use S-Parameters for Transistor Circuit Design", EEE, December, 1965, p. 36.

33. Norris, P.M., "Using Linvill Techniques for R.F. Amplifiers", Motorola Semiconductor Products Inc., Application Note #AN-166.

34. Wollesen, D.L., "UHF Transmission-Line Power Amplifier Designed with Smith Chart Techniques", Motorola Semiconductor Products Inc., Application Note #AN-217.

35. Mertens, D.E., "A New Look at the Parallel Tuned Circuit", Electro-Technology, May, 1964, pp. 58-62.

36. Lauchner, J., and Silverstein, M., "Transistor Amplifiers Using the Linvill Technique", Motorola Engineering Bulletin, Vol. 13, #4, 1965; Also: Electronic Design (condensed version), Vol. 14, #8, April 12, 1966, pp. 47-54.

37. Hetterscheid, W. Th., "Transistor Bandpass Amplifiers", Phillips Publishing Co., Eindhoven, The Netherlands, 1964.

38. Norris, P.M., "Small-Signal Amplifier Design Using Admittance Parameters", EDN, October, 1965, p. 130.

39. Martes, G.E., "Make Impedance Matching Easier", Electronic Design, July 5, 1966, pp. 46-48.

40. Blanchard, W.C., "New and More Accurate Inside-Out Smith Chart", MicroWaves, November, 1967, pp. 38-42.

41. Smith, P.H., "Transmission Line Calculator", Electronics, January, 1939, pp. 29-31.

42. Campbell, D., "Smith-Chart Short Cut Renormalizes Feedline I Impedances", Electronics, November 16, 1962, pp. 53-54.

43. Smith, P.H., "An Improved Transmission Line Calculator", Electronics, January, 1944, pp. 130-133, 318-325.

44. Dawirs, H.N., "A Chart for Analyzing Transmission-Line Filters From, Input Impedance Characteristics", Proc. of IRE, April, 1955, pp. 436-443.

45. Panisello, A.A., "Pi-Network as East as Pie", EDN, July, 1967, p. 58.

46. Geckle, Jr., W.A., "Optimum Pi Pad Design", EEE, September, 1965, pp. 62, 63.

47. Matthaei, G.L., "Synthesis of Tchehycheff Impedance-Matching Networks, Filters, and Interstages", IRE Trans. on Circuit Theory, Vol. CT-3, September, 1956, pp. 163-172.

48. Anderson, A.E., "How to Measure Source Impedance with a Slotted Line", EEE, November, 1965, p. 52.

49. Gardiol, F.E., "Chebyshev Transformer Nomograms", MicroWaves, November, 1967, pp. 56-58.

50. Hall, A.H., "Impedance Matching by Tapered or Stepped Transmission Lines", The Microwave Journal, March, 1966, pp. 109-114.

51. Dougharty, W., "The Theory of Broadband Coupled Circuits", Electronic Engineering ((British)), August, 1959, pp. 488-494.

52. Mueller, W.P., and Overmier, E.E., "Designing π Networks for Desired Q", Electronic Products, September, 1964, p. 19.

53. WcWhorter, M.M., and Pettit, J.M., "The Design of Stagger-Tuned Double-Tuned Amplifiers for Arbitrarily Large Bandwidth", Proc. of IRE, August, 1955, pp. 923-931, and, Proc. of IRE, Nov., 1955, p. 1579.

54. Dishall, M., "Exact Design and Analysis of Double-and Triple-Tuned Band-Pass Amplifiers", Proc. of IRE, June, 1947, pp. 606-626.

55. Levy, G.R., "Design Considerations for the RCA 2N2102 in Low-to-High Level, and DC-to-RF Applications", RCA Application Note #SMA-14.

56. Vincent, B.T., "Large Signal Operation of Microwave Transistors", IEEE Trans. on MTT, (Correspondence), Vol, MTT-13, No. 6, November, 1965, pp. 865-866.

57. Bauhaus, R.H., and Wong, R.W., "Predicting Transistor Behavior at the Threshold of Large Signal Conditions", MicroWaves, October, 1969, p. 40.

58. Müller, O., "Large Signal S-Parameter Measurements of Class-C Operated Transistors", NTZ-Communication Journal, Vol. 21, No. 10, October, 1968, pp. 644-647

59. "Transistor Parameter Measurements", Hewlett-Packard Application Note 77-1.

60. Roos, E., "Graphical Approach to Networks Challenged", Electronic Design ((Letters)), April 25, 1968, pp. 48-50.

61. Johnson, R., "Draw Your Own Networks' Impedance", Electronic Design, January 4, 1968, pp. 102, 103.

62. Terman, F.E., Radio Engineers' Handbook, McGraw Hill, 1943, pp. 85, 93, 109, 119, 135-148, 210-214.

63. Joyce, M.V., and Clarke, K.K., "Transistor Circuit Analysis", Addison-Wesley Pub. Co., Inc. 1961, pp. 247-263.

64. Walston, J.A., and Miller, J.R., "Transistor Circuit Design", McGraw Hill, 1963, pp. 345-359.

65. Jasik, H., Antenna Engineering Handbook, McGraw Hill, First Edition, 1961, Chap. 35, pp. 17, 18.

67. Linvill, J.G., and Gibbons, J.F., "Transistor and Active Circuits", McGraw Hill, 1961, Charts 11.1 to 11.3, and 14.1 and 14.2.

68. Magil,L.M., "Watt-Megahertz Ratings Run Second to Reliability in Foreign RF Power Transistors", Electronics, April 27, 1970, pp. 80-89.

69. The Microwave Engineers' Handbook and Buyer's Guide - 1970, Horizon House Publishers, pp. 102-103.

70. Gundlach, R., "Rx for RF Power Transistors", Electronics, May 26, 1969, pp. 84-90.

71. Chao, S.C., "Computer-Aided Design of Broadband UHF Transistor Amplifiers", The Microwave Journal, July, 1969, pp. 79-83.

72. Lee, H.C., "Microwave Power Transistors", The Microwave Journal, February, 1969, pp. 51-65.

73. Carley, D.R., "Transistor Amplifiers", The Microwave Journal, July, 1969, pp. 52-56.

74. Ray, C.H., and Cavin, R.K., III, "The Q of the Two-Branch RLC Circuit", IEEE Trans. on Education, March, 1969, pp. 52, 53.

75. Cowdell, R.B., "Universal Graphs Map Direct Route to Mismatched Pi and Tee Filters", Electronics, November, 1969, pp. 121-128.

76. Belohoubek, et.al., "Hybrid Integrated 10-Watt CW Broadband Power Source at S Band", IEEE Journal of Solid State Circuits, Vol. SC-4, No. 6, December, 1969, pp. 360-366.

77. Cooke, H.,"An Inside Look Into Microwave Transistors Part 1: Power Devices", MicroWaves, December, 1969, pp. 46-52.

78. Bailey, R.L., "Large-Signal Nonlinear Analysis of a High-Power High-Frequency Junction Transistor", IEEE Trans. on Electron Devices, Vol. ED-17, No. 2, February, 1970, pp. 108-119.

79. Flynn, G., "Forum on High-Frequency (S-Band) High-Power Transistors", Electronic Products, February, 1970, pp. 27-35.

80. Tatum, J.G., "Microwave Transistor-Parameter Trade-Offs in Circuit Design, Part 1: Basic Considerations", MicroWaves, September, 1967, pp. 26-32.

CUSTOMER APPLICATION NOTES, DATA SHEETS, AND TECHNICAL ARTICLES RECOMMENDING ATC CAPACITORS

RELEASE DATE	RECOMMENDING CUSTOMER	APPLIC. NOTE #	DATA SHEET #	ARTICLE (OR) REFERENCE
18 Jan 69	ITT Semiconductors			"Build Broadband RF Amplifiers" by J.A. Benjamin, Electronic Design, Jan.18, 1969, pp.50-54
———	ITT Semiconductors		3TE609	50.Watt, 400 MHz, Power Transistor
Nov. 69	Motorola Semiconductors		MM4019	Complementary NPN/PNP VHF Amplifie
Feb. 70	Bell Telephone Laboratories			"Methods of Reducing Noise of Junction Field Effect Transistor (JFET) Amplifiers" by H.E. Kern and J.M. McKenzie, IEEE Transactions on Nuclear Sciences, Feb. 70, pp. 260-261
3-70	RCA Electronic Components	427	2N5921	L-Band Power Amplifier/Transistor
———	Kertron, Inc.		3TX820 3TX821	2.Watt, 470. MHz Transistors
3-70	RCA Solid State Division	426	2N5919	16.Watt, 225-400 MHz Amplifier Transistor
3 June 70	Bell Telephone Laboratories			"Lumped Constant, Hard Substrate, High Power Phase Shifters", by Peter Onno and Andrew Plitkins (Paper delivered at the MIC Seminar at Monmouth College, Fort Monmouth, N.J. on 3 June,1970)
7-70	RCA Solid State Division	448	2N5918	10.Watt, 225 to 400 MHz Amplifier Transistor
———	Kertron, Inc.		3TX632	5.Watt, 470. MHz, Power Transistor
July 71	Wheeler Labs			"An L-Band MIC Front End for an IFF Receiver", by R. Giannini, S. Anghel, and R. Camisa, IEEE Trans. on MTT, Vol. MTT-19, No. 7, July, 1971, pages 622-627.
9-71	R.C.A. Solid State Division		TA7995	10.Watt, 2. GHz, Emitter-Ballasted Silicon NPN Overlay Transistor

CUSTOMER APPLICATION NOTES, DATA SHEETS, ND TECHNICAL ARTICLES RECOMMENDING ATC CAPACITORS

ELEASE TE	RECOMMENDING CUSTOMER	APPLIC. NOTE	DATA SHEET #	ARTICLE (OR) REFERENCE
7-70	RCA Solid State Division		TA7706	25 Watt, 225 to 400 MHz Amplifier Transistors.
10-70	RCA Solid State Division	AN4421		16-and 25-Watt Broadband Power Amplifiers using RCA-2N5918, 2N5919, and TA 7706 UHF/Micro-wave Power Transistors.
1-71	RCA Solid State Division		TA7994	5 Watt, 2 GHz, Emitter-Ballasted Silicon N-P-N Overlay Transistor.
1-71	Motorola Semiconductors	DS5460	2N5923	2.5 Watt, 1.0 GHz, NPN, Silicon Power Transistor
1-71	Motorola Semiconductors	DS5461	2N5924	5.0 Watt, 1.0 GHz, NPN,Silicon RF Power Transistor
1-71	Motorola Semiconductors	DS5462	2N5925	10.Watt, 1.0 GHz, NPN, Silicon RF Power Transistor
12-70	Communications Transistor Corp.	2.1.8.3C	B70-12	70.Watt, 175. MHz., 12. volt Land Mobile RF Power Transistor
ⁿ-70	Communications Transistor Corp.	2.1.8.4C		175.MHz., 12.volt, 140.Watt, RF Power Amplifier
3-71	Communications Transistor Corp.	2.2.8.7B	C40-28	400. MHz, 28. Volt, RF Power Transistor
-71	Solitron Devices, Inc.			"40 Watt Broadband VHF Power Amplifier", Microwave Systems News, May/June, 1971, p. 28
9-70	Solid State Scientific Inc.		2N5589	3 Watt silicon NPN VHF Communications Transistor
-70	Solid State Scientific Inc.		2N5590	10.Watt silicon NPN VHF Communications Transistor
-70	Solid State Scientific Inc.		2N5591	25.Watt silicon NPN VHF Communications Transistor

CUSTOMERS RECOMMENDING ATC

CUSTOMER APPLICATION NOTES, DATA SHEETS, AND TECHNICAL ARTICLES RECOMMENDING ATC CAPACITORS

RELEASE DATE	RECOMMENDING CUSTOMER	APPLIC. NOTE #	DATA SHEET #	ARTICLE (OR) REFERENCE
2-71	FAIRCHILD Microwave and Optoelectronics Div.		MT1412	12 Watts, 470. MHz, 12.5 VDC NPN large signal UHF power transistor
2-71	FAIRCHILD Microwave and Optoelectronics Div.		MT 1425	25. Watts, 470. MHz, 12.5 VDC NPN large signal UHF power transistor
—	Solitron Devices		C-A-1186	2 stage XMTR, 13V, 175 MHz, 28 Watts using SRK-2044 devices
5-71	Bell Laboratories Whippany, N.J.			"Miniature Multi-Kilowatt PIN Diode MIC Digital Phase Shifters", by Peter Onno and Andrew Plitkins, IEEE-GMTT International Microwave Symposium Proceedings, May 17, 1971, page 22
—	Solitron Devices Inc.		C-B-1190	R.F. Power Amp, 3 stage 150-175 MHz
—	Solitron Devices Inc.		C-A-1180	3 Stage XMTR, 600 MHz, 24 V.
9-71	Motorola Semiconductor Products Inc	AN-548		25. Watt 450.-512 MHz 12.5 VDC RF Amplifier, using 2N5945, 2N5946, and 2N6136.
—	Communications Transistor Corp.	—	—	"A 140-W, 175-MHz Amplifier: Where Low-Loss Capacitors Are a Must" (by R. Cushman) EDN/EEE, Nov. 15, 1971, page 34.
6-70	RCA Solid State Division	File #440	2N5920 (TA7487)	Silicon NPN overlay coaxial transistor, 3. Watts, 12. dB gain at 1. GHz and 28. VDC.

CUSTOMER APPLICATION NOTES, DATA SHEETS,
ND TECHNICAL ARTICLES RECOMMENDING ATC CAPACITORS

EASE TE	RECOMMENDING CUSTOMER	APPLIC. NOTE #	DATA SHEET #	ARTICLE (OR) REFERENCE
. 1971	Communications Transistor Corp.	2.0.8DA	CT XB50-28	Wideband 28. Volt High gain 50.Watt UHF Power Transistor
-71	RCA Solid State Division	File#504	2N6104 2N6105	30-W, 400-MHz Broadband Emitter-Ballasted Silicon N-P-N Overlay Transistors
-71	RCA Solid State Division	File#543	2N6265	2-W, 2-GHz, Emitter Ballasted Silicon N-P-N Overlay Transistor
-71	RCA Solid State Division	File#544	2N6266	5-W, 2-GHz, Emitter Ballasted Silicon N-P-N Overlay Transistor
-71	RCA Solid State Division	MPT-700		RCA Microwave Power Transistors Design Features; Circuit Applications
-71	RCA Solid State Division	File#538	40898 40899	6- and 2-W, 2.3 GHz Emitter Ballasted Silicon N-P-N Overlay Transistors
-7	RCA Solid State Division	File#546	2N6268 2N6269	6.5-and 2-W, 2.3 GHz Emitter-Ballasted Silicon N-P-N Overlay Transistors
-7	RCA Solid State Division	File#545	2N6267	10-W, 2-GHz, Emitter Ballasted Silicon N-P-N Overlay Transistors
b. 1972	Communications Transistor Corp.	File # 2.3.8.3A	E1-28	1. Watt 2 GHz 28 Volt Power Transistor
b. 1972	Communications Transistor Corp.	File # 2.3.8.3A	E3-28	2.5 Watt 2 GHz 28 Volt Power Transistor
b. 1972	Communications Transistor Corp.	File # 2.3.8.3A	ES-28	5. Watt 2 GHz 28 Volt Power Transistor
b. 1972	Communications Transistor Corp.	File # 2.3.8.3A	E10-28	10. Watt 2 GHz 28 Volt Power Transistor

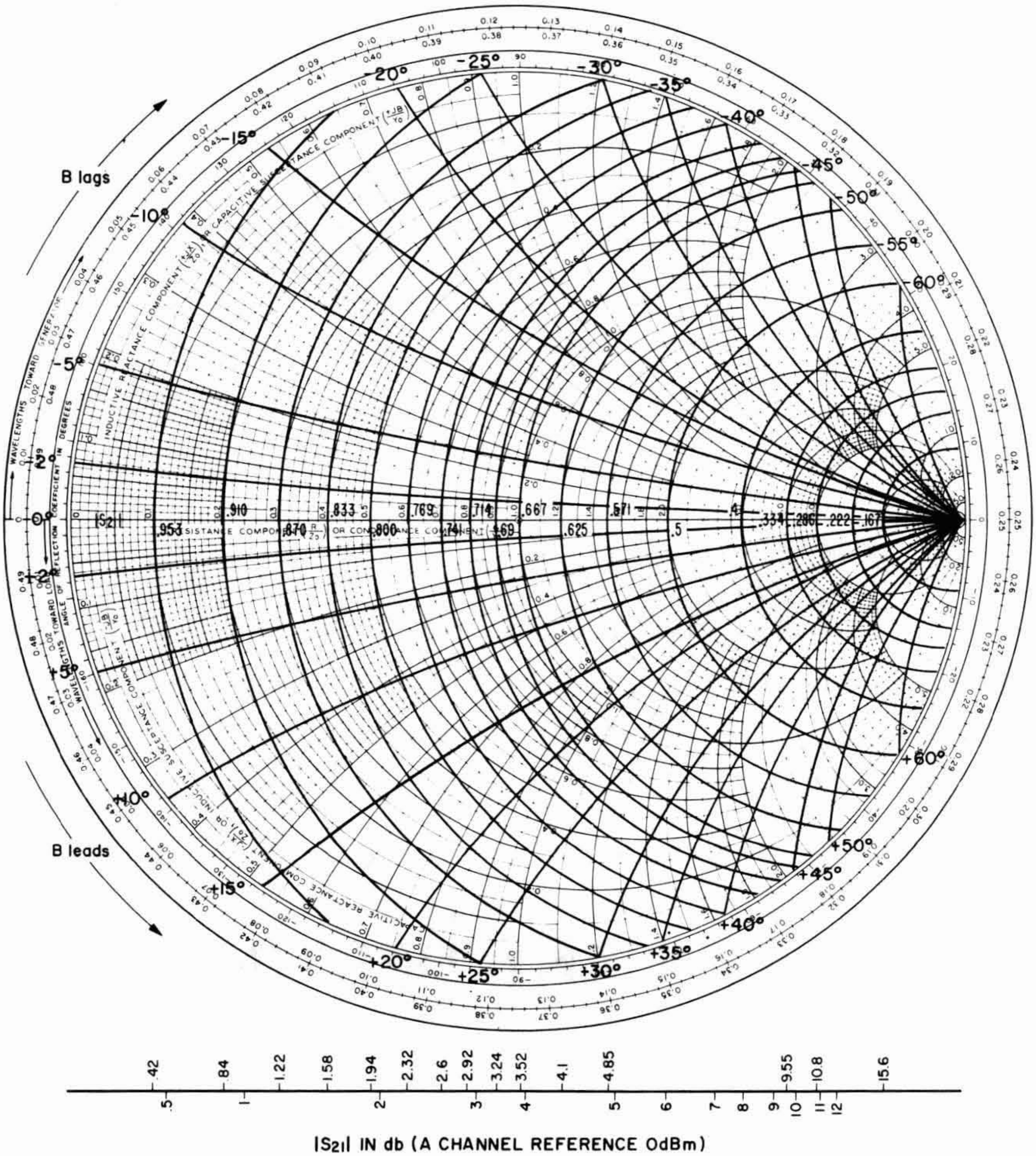
CUSTOMER APPLICATION NOTES, DATA SHEETS, AND TECHNICAL ARTICLES RECOMMENDING ATC CAPACITORS

RELEASE DATE	RECOMMENDING CUSTOMER	APPLIC. NOTE#	DATA SHEET#	ARTICLE (OR) REFERENCE
8-71	RCA Solid State Division	File # 505	2N5919A	16-W, 400-MHz, Silicon N Emitter-Ballasted Overla Transistor
11-71	RCA Solid State Division	An-4774		Hot Spotting in RF Powe Transistors
8-74	Hewlett-Packard	Appli. Note 949-1		"Linear Power Amplificat Using the H.P. 35850 Se Transistors"

SECTION 6

SUPPLEMENTARY DESIGN AIDS

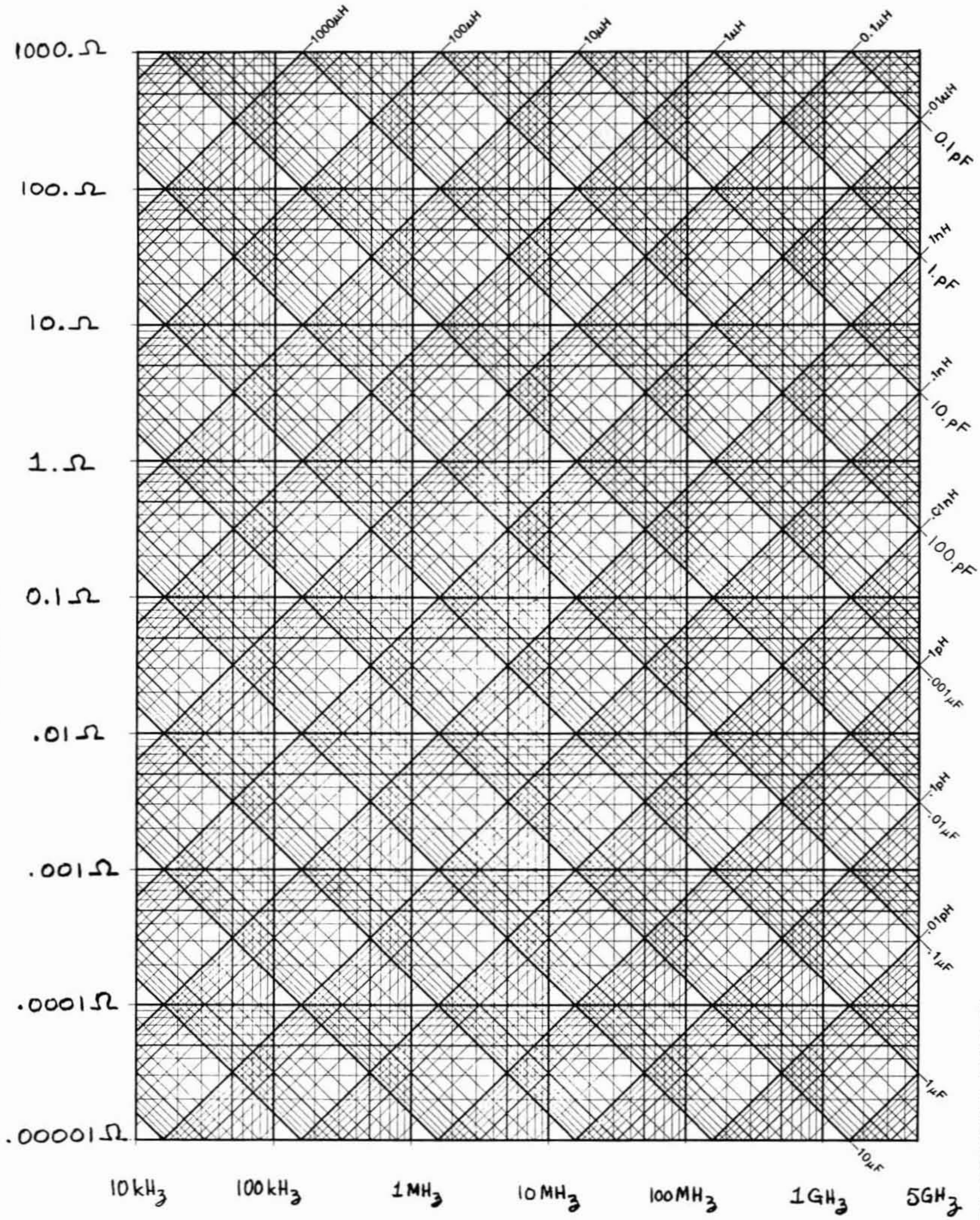
Reprinted with permission from: PRD Electronics, Inc.
Westbury, New York
Application Note #22 on: VECTOR VOLTMETER MEASUREMENT TECHNIQUES



(Series) S-Parameter ◄──► Impedance

Conversion Chart

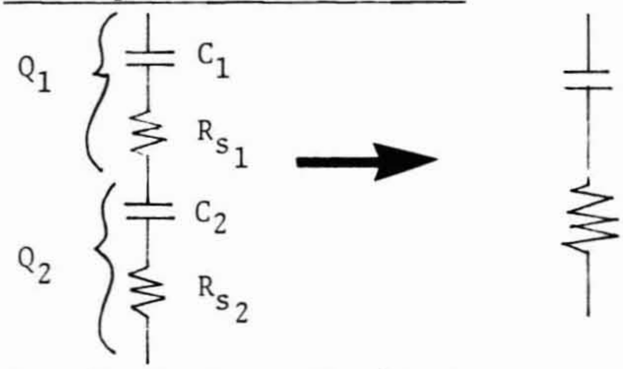
CAPACITOR OR INDUCTOR REACTANCE VS. FREQUENCY



FREQUENCY

THE EFFECTIVE Q_{cap} OF CAPACITOR COMBINATIONS

(I) Two capacitors in series:



$$C_{eff} = \left(\frac{C_1 \times C_2}{C_1 + C_2}\right)$$

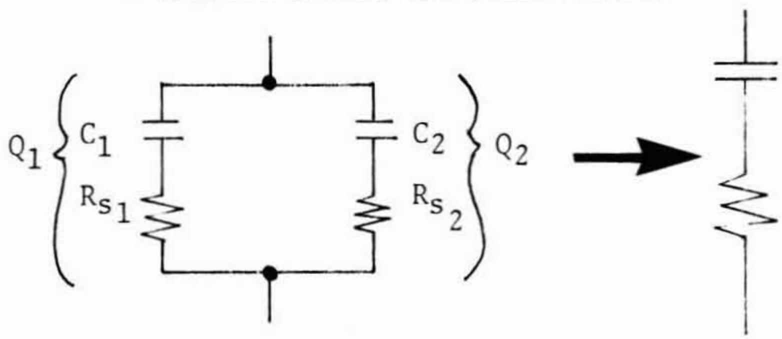
$$R_{eff} = (R_{s_1} + R_{s_2})$$

$Q_1 = (X_{c_1}/R_{s_1}) \quad Q_2 = (X_{c_2}/R_{s_2})$

$$Q_{eff} = \left[\frac{Q_1\ Q_2\ (C_1 + C_2)}{Q_1\ C_2 + Q_2\ C_1}\right]$$

Or, otherwise stated: $Q_{eff} = \left(\frac{X_{c_{eff}}}{R_{eff}}\right)$

(II) Two capacitors in parallel:



$$C_{eff} = (C_1 + C_2)$$

$$R_{eff} = \left(\frac{R_{s_1} \times R_{s_2}}{R_{s_1} + R_{s_2}}\right)$$

Q values: (a) For any values of capacitance having equal Q's, the effective Q_{cap} of the combined pair equals that of either of the individual capacitors above.

(b) For equal values of capacitance but with unequal Q's, the effective Q_{cap} of the combined pair is given by:

$$Q_{eff} = 2\left(\frac{Q_1 \times Q_2}{Q_1 + Q_2}\right)$$

(c) For the case of both unequal capacitance values and unequal Q_{cap} values:

$$Q_{eff} = \left(\frac{X_{c_{eff}}}{R_{eff}}\right), \text{ or, alternatively: } Q_{eff} = \left[\frac{Q_1\ Q_2\ (C_1 + C_2)}{C_1\ Q_2 + C_2\ Q_1}\right]$$

For much of the above, see:
The Handbook of Chemistry and Physics, The Chemical Rubber Co., 41st. Edition.

COMPUTER-AIDED RF CIRCUIT DESIGN SERVICES

SOFTWARE PACKAGE NAME	COMPANY OFFERING SERVICE
C.O.D. and M.O.D.	Optimal Systems Research, Inc. P.O. Box 1169 Wall, N.J. 07719
FILTAN and FILTEST	Sanders Associates, Inc. Microwave Division P.O. Box 907 Nashua, N.H. 03060
AMCAP, FILTAN, MIFIL, ASWMDC, QWT, COLPAF, SYMCO, ASYMCO	Environmental Computing 21 George Street Lowell, Mass. 01852
MATCH and MICRONET	Applicon, Inc. 83 Second Avenue Burlington, Mass. 01803
AMCAP, MIFIL, FILTER, ACAP	On-Line Systems 230 Park Avenue New York, N.Y. 10017
MICAP	Tymshare Inc. 464 Hudson Terrace Englewood Cliffs, N.J. 07647
MICAP	Tymshare Inc. University Avenue Palo Alto, California
AMCAP, ACAP, FILTAN, IMPACT	Tymshare Inc. Bubb Road Cupertino, California
SCEPTRE	Control Data Corp. 60 Hickory Drive Waltham, Mass. 02154
MCAP	Shared Applications, Inc. 213 E. Washington Street Ann Arbor, Michigan 58108
ANCIR/TS	W.W. Gaertner Research, Inc. 205 Saddle Hill Road Stamford, Conn. 06903
OPTINET	Dean Hall Assoc., Inc. 200 Third Street Los Altos, California 94022

COMPUTER-AIDED RF CIRCUIT DESIGN SERVICES

SOFTWARE PACKAGE NAME	COMPANY OFFERING SERVICE
DEMON	Electronic Components Laboratory U.S. Army Electronics Command Fort Monmouth, N.J.
SCAMAT	Instituto di Elettronica Universita di Bologna Italy
AMCAP, MATCH 1, LPFILT, ACTFIL	On-Line Systems, Inc. 425 Broad Hollow Road Melville, N.Y. 11746
MATCH	Com-Share Inc. P.O. Box 1588 Ann Arbor, Mich.
MATCH	Applied Logic Corp. P.O. Box 124 One Palmer Square Princeton, N.J.
MATCH	Multicomp Inc. 36 Washington St. Wellesley Hills, Mass.
AMCAP, ACAP FILTAN, IMPACT	On-Line Systems Inc. 40 Washington St. Wellesley, Mass. 02181
IMPACT and AMCAP	On-Line Systems McKnight Road Pittsburg, Penna.
AMCAP, ACAP, MIFIL FILTAN, IMPACT	Comp/Utility Inc. One Center Plaza Boston, Mass. 02108
FILTAN	Compu-Time 327 Orange Avenue Daytona Beach, Fla. 32014
AMCAP and IMPACT	Data Line Ltd. Saint Clair West Toronto, Ontario, Canada
OPTINET	National CSS, Inc. 460 Summer Street Stamford, Conn. 06891

NATIONAL CSS, INC.

Corporate Headquarters:

460 Summer Street
Stamford, Connecticut 06901
(203) 327-9100

Branch Offices:

BETHLEHEM
800 Ostrum Street
Bethlehem, Pennsylvania 18015
(215) 691-3616

CAMBRIDGE
1033 Massachusetts Avenue
Cambridge, Massachusetts 02138
(617) 868-2950

CHICAGO
333 North Michigan Avénue
Chicago, Illinois 60601
(312) 368-8770

ELIZABETH
27 Prince Street
Elizabeth, New Jersey 07208
(201) 965-2250

LOS ANGELES
4311 Wilshire Boulevard
Los Angeles, California 90005
(213) 937-4600

Branch Offices:

SAN FRANCISCO
One California Street
San Francisco, California 94101
(415) 989-3930

SUNNYVALE
430 South Pastoria Avenue
Sunnyvale, California 94086
(408) 739-6271

NEW YORK
485 Madison Avenue
New York, New York 10022
(212) 688-7930

PHILADELPHIA
Three Pen Center
Philadelphia, Pennsylvania 19102
(215) 665-1566

ROCHESTER
110 Allens Creek Road
Rochester, New York 14618
(716) 442-7660

For further information on "Who offers What in Remote Access", contact:

Janet M. Taplin or Eugene C. Gaines, Jr.
Time-Sharing Enterprises, Inc.
University City Science Center, Bldg. 1.
3401 Market Street
Philadelphia, Pa. 19104
Tel: (215) 386-6250

DESIGN TRICKS:

HOW TO MAKE LOW-LOSS VHF INDUCTORS

Extracts and Quotes
(from Solitron Devices, Inc. Application Note of 26 April 1971 on a 40. Watt 175. MHz RF Amplifier operating at 12.5 VDC)

Instead of having to make a special purchase of silver metal ribbon to manufacture high Q inductors, "American Technical Ceramics capacitors" may additionally be employed "as inductors, as well as DC-blocks, with their flat leads cut to specific lengths. The ATC capacitors have good low-loss characteristics above 100. MHz."

DESIGN TRICKS:

HOW TO DEVISE A LOW-LOSS INDUCTOR FOR S-BAND

Even capacitors without leads exhibit the phenomenon of series-resonance if employed at a sufficiently high radio-frequency. This is due to the fact that they have a finite physical length plus some inescapable residual inductance inherent in manufacturing processes. Furthermore, the larger the capacitance value, the lower the frequency at which this resonance will occur.

If the capacitor is carefully constructed, this inductance will be both extremely small (sub-nanohenry) and consistently reproducible. In such units, this normally undesirable characteristic may be turned to advantage, if the basic design of the capacitor also incorporates a dielectric material having extremely high Q.

A growing number of ATC's customers have been employing high-valued ATC 100-B capacitors at several times their commonly utilized range (below 1000. MHz) as inductive elements of impedance matching networks. Used in this manner, 470. pF and 1000. pF units provide a further unusual combination of advantages:

- a microwave inductor that is also low-loss,
- having a built-in DC-block,
- easy to handle mechanically,
- low in cost,
- providing considerable design flexibility for attempting novel subminiature circuitry.