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THE PERFORMANCE OF 10 μF POWER LINE FEEDTHROUGH CAPACITORS

The Moore School of Electrical Engineering
University of Pennsylvania

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THE PERFORMANCE OF 10 μF POWER LINE FEEDTHROUGH CAPACITORS

Donald E. Groff
The Moore School of Electrical Engineering
University of Pennsylvania

FOREWORD

This technical report was prepared by Mr. Donald E. Groff, The Moore School of Electrical Engineering, University of Pennsylvania, under Contract F30602-68-C-0178, Project 4540, Task 454001. Secondary report number is 69-19.

Mr. Robert L. White (EMCVM-1) was the Rome Air Development Center project engineer.

The distribution of this report is limited because the document contains the performance characteristics of a test instrument which is used for specification testing of military equipments.

This technical report has been reviewed and is approved.

Approved: Robert L. White
ROBERT L. WHITE
Effort Engineer

Approved: Samuel H. Zaccari
for RICHARD M. COSEL, Colonel, USAF
Chief, Communications Division

FOR THE COMMANDER: Irving J. Gabelman
IRVING J. GABELMAN
Chief, Advanced Studies Group

ABSTRACT

The insertion loss of a number of power line feedthrough filter capacitors shows pronounced deviation from the expected results in the HF region, which is traced to the low impedance transmission line character of the filter.

The insertion loss of the filter is predicted from models of varying degrees of sophistication. The validity of considering a rolled foil capacitor as a transmission line is discussed, the theory indicating a series of dips in the insertion loss, and a series of poles and zeroes in the input impedance.

Experimental results clearly indicate transmission line effects in a number of feedthrough units. Similar behavior was also detected in conventional rolled foil capacitors. The practical significance of the results is discussed along with the use of such filters for measuring spurious emission from electronic equipment.

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THE PERFORMANCE OF 10 μF POWER LINE FEEDTHROUGH CAPACITORS

1.0 INTRODUCTION

This report discusses the performance of power line feedthrough capacitors at radio frequencies, in connection with their use in electromagnetic compatibility testing.

Such capacitors are used in measuring spurious power line emissions from electronic equipment. Putting the power lines through these capacitors serves two purposes: (a) to prevent spurious signals on the power line from affecting the measurement, and (b) to standardize the impedance seen from the equipment power terminals. The effectiveness of the capacitor will depend, among other things, on the RF impedance of the power line, both looking out toward the power source, and in toward the equipment.

1.1 EMISSION TEST CIRCUIT

The conducted emission test in military standard MIL-STD-462 calls for placing the equipment on a ground plane and passing the two power leads through 10 μF feedthrough capacitors at some distance from the unit under test [1]. This approximates the effect of a short circuit at radio frequencies. The RF currents at points a and b on Fig. 1 are measured with a current probe. The RF circuit is approximately that of Fig. 2, for which the current measurement yields the value of $\left|\frac{V}{Z}\right|$. (Actually for high accuracy a 3-terminal equivalent network should be considered, because at radio frequencies the ground plane carries significant currents.) The value of Z, the power terminal RF output impedance, is important. It should be at least 10 times the impedance of the feedthrough capacitor if the short circuit output approximation is to be valid.

1.2 PERFORMANCE OF 10 μF UNITS

The 10 μF feedthrough capacitor has a theoretical reactance of 260 ohms at 60 Hz, 1.6 ohms at 10 kHz, .016 ohms at 1 MHz, etc. Hence, one might expect it to act as a very effective

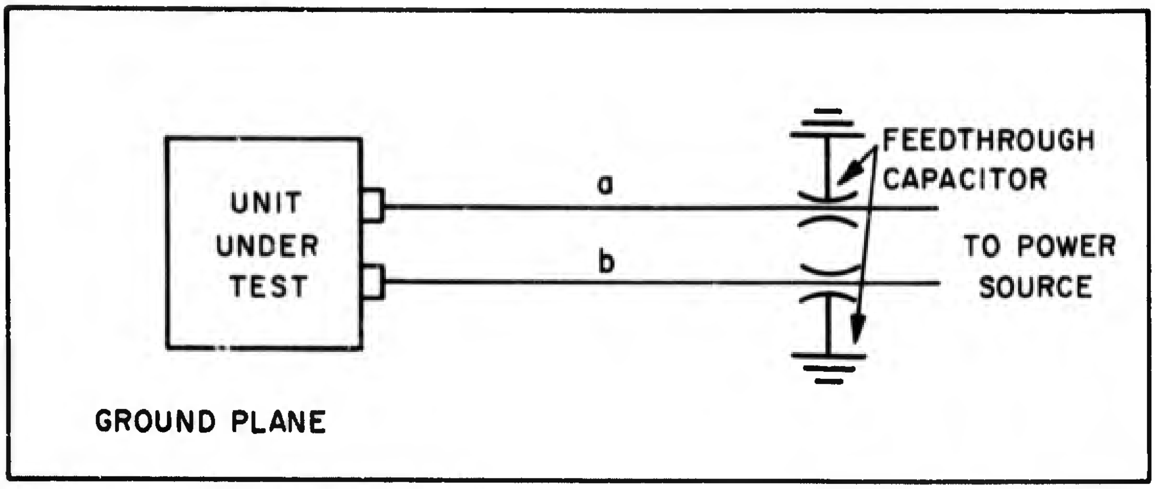

Figure 1 SETUP FOR MEASURING POWER LINE EMISSION

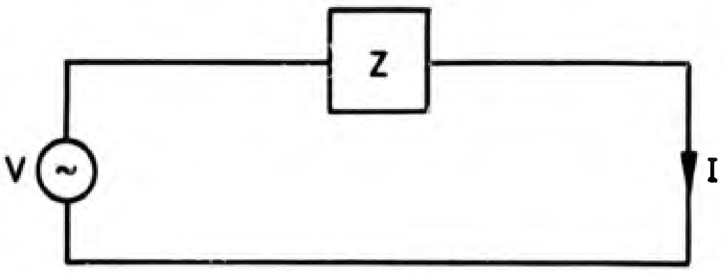

Figure 2 SIMPLIFIED EQUIVALENT CIRCUIT FOR FIGURE 1

short circuit, especially since many pieces of equipment have a simple .01 μF output filter. However, one must also consider other impedances in the circuit such as the stray lead inductance. For example, 0.1 μH yields series resonance at about 150 kHz. The amount of stray inductance may be estimated by the formula for the inductance of a straight wire [2]

$$L = \frac{\mu}{4\pi} \ell \left[\ln \frac{2\ell}{r} - 1 \right] \tag{1}$$

where μ is the permeability, ℓ the length, and r the radius of the wire. For example, if ℓ = 25 cm = .25 m, and r = 2 mm = .002 m, L = 0.11 μH. The performance of rolled-foil capacitors at radio frequencies is not a widely discussed topic, except for the matter of the series inductance. Several references indicate that such units are usable to about 1 MHz [3, 4]. SAE document ARP-936, which is concerned with feedthrough units [5], requires an insertion loss which does not increase above 150 kHz. Another reference notes deviations from expected behavior, but does not elaborate [6].

An extensive series of tests was therefore performed on the 10 μF feedthrough capacitors, and on several other rolled-foil capacitors, of both feedthrough and ordinary construction. In the 1-5 MHz range, most of the feedthrough units showed pronounced deviation from ideal behavior. Tests on several conventional units showed similar, but much less clear, deviations. The feedthrough units had resonances in insertion loss and input impedance, and most importantly, showed significant delay between input and output terminals. This immediately suggested transmission line behavior. It should be noted that in effect a feedthrough capacitor resembles a section of low impedance line.

2.0 THEORY OF MEASUREMENTS

Measuring insertion loss is a conventional way of determining the performance of transmission line filters [7]. It is also a convenient method for measuring the magnitude of small impedances at radio frequencies.

Insertion loss α may be defined as the ratio:

$$\frac{\text{power delivered to a matched load without the filter}}{\text{power delivered to a matched load with the filter}} ,$$

as measured in the circuit of Fig. 3.

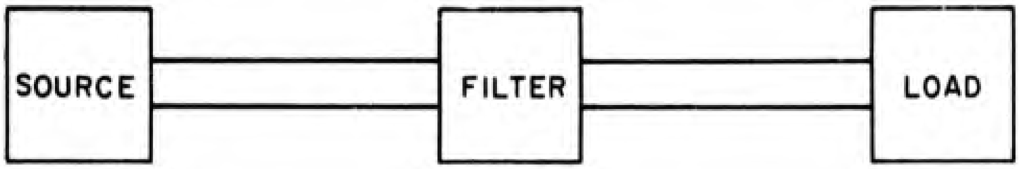

Figure 3 MEASUREMENT OF INSERTION LOSS

Considering the T equivalent of a two-part filter connected with coaxial lines of characteristic impedance R ohms, we have the equivalent circuit shown in Fig. 4.

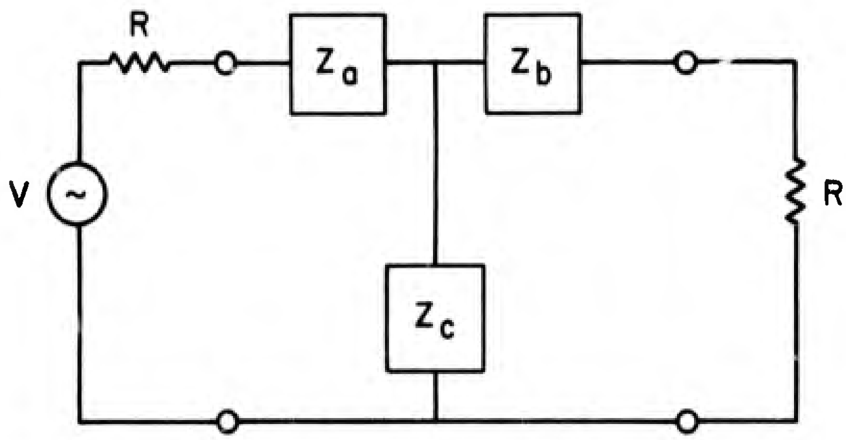

Figure 4 TWO PORT T EQUIVALENT CIRCUIT

The power delivered to the load without the filter is:

$$P = \frac{V^2}{4R}$$

and with the filter, the result is:

$$P = \frac{V^2}{R}\left|\frac{Z_c}{Z_a + Z_b + Z_c + R + \dfrac{Z_aZ_b + Z_bZ_c + Z_aZ_c}{R}}\right|^2$$

so that the insertion loss α is:

$$\alpha = \left| \frac{Z_a + Z_b + 2Z_c + R + \dfrac{Z_aZ_b + Z_bZ_c + Z_aZ_c}{R}}{2Z_c} \right|^2 \tag{2}$$

2.1 IDEAL CAPACITOR

For an ideal capacitor, with negligible lead inductance, $Z_a = Z_b = 0$ and $Z_c = \frac{1}{j\omega C}$, so eq. (1) reduces to

$$\alpha = 1 + \frac{1}{4}\omega^2 R^2 C^2 \tag{3}$$

For $\omega \gg \frac{2}{RC}$, $\alpha \propto f^2$.

2.2 IDEAL INDUCTOR

If the shunt element Z_c is an inductive reactance, $j\omega L$, and $Z_a = 0 = Z_b$, then

$$\alpha = 1 + \frac{R^2}{4\omega^2 L^2} \tag{4}$$

For $\omega \ll \frac{R}{2L}$, $\alpha \propto \frac{1}{f^2}$. This would apply to a conventional capacitor well above the series resonance frequency.

2.3 IDEAL CAPACITOR WITH LEAD INDUCTANCE

If we include lead inductance, and let $Z_a = Z_b = j\omega L$, with $Z_c = \frac{1}{j\omega C}$, the expression reduces to

$$\alpha = (1 - \omega^2 LC)^2 + \frac{\omega^2 R^2 C^2}{4} \left(1 - \frac{\omega^2 L^2}{R^2} + 2\frac{L}{R^2 C}\right)^2 \tag{5}$$

The minimum value of L is not likely to be greater than 0.1 μH.

For

$$L = 10^{-7}\ H$$

$$C = 10^{-5}\ F$$

$$R = 50\ \Omega$$

we have

$$\alpha = (1 - \omega^2\ LC)^2 + \frac{\omega^2 R^2 C^2}{4}\left(1 - \omega^2 4 \times 10^{-18} + 4 \times 10^{-6}\right)^2$$

and see that the second bracketed term is essentially unity for $\omega < 10^8\ \sec^{-1}$ or 15 MHz. This leads to

$$\alpha = 1 + \omega^2\left(\frac{R^2C^2}{4} - 2LC\right) + \omega^4\ L^2\ C^2\ .$$

Numerically

$$\alpha = 1 + \omega^2(6 \times 10^{-5} - 2 \times 10^{-12}) + \omega^4\ 10^{-24}$$

so that the significant terms are

$$\alpha = 1 + \omega^2\ \frac{R^2C^2}{4} + \omega^4 L^2 C^2\ \ ,$$

and the ω^4 predominates only above 40 MHz. For this example, the indication is that the insertion loss is essentially unaffected by lead inductance below 15 MHz. It would, of course, affect the input impedance, producing a series of resonance.

2.4 TRANSMISSION LINE MODEL

If the filter consists of a low impedance transmission line segment, the results are a good deal different. The analysis, given in detail in the Appendix, is summarized here.

A length ℓ of transmission line of low characteristic impedance Z_o (Z_o << R) has a time delay $\tau = \ell/v$, where v is the speed of wave propagation on the line. For low frequencies, this line acts like a capacitance $C = \frac{v}{\ell Z_o}$. The insertion loss of this filter is given by (see Appendix for derivation)

$$\alpha = 1 + \frac{1}{4}\frac{R^2}{Z_o^2}\sin^2 \omega\frac{\ell}{v} \quad , \qquad (6)$$

and the open circuit input impedance is given by $-jZ_o \cot \omega \frac{\ell}{v}$. As ω is varied, the insertion loss has a maximum value of $1 + \frac{1}{4}\frac{R^2}{Z_o^2}$, and periodically drops to unity, and the input impedance has a series of poles and zeroes. This ignores the external lead inductance, and assumes the line to be uniform and lossless.

The characteristic impedance may be estimated from the dimensions of the structure, which indicates values of the order of 0.1 ohm. There are also three ways to measure Z_o:

1. The geometric mean of the short-circuited and open-circuited input impedances,

$$\sqrt{Z_{oc}\, Z_{sc}} \quad , \qquad (7)$$

is equal to Z_o. The measurement of these impedances must be done at a low enough frequency for lead inductance to be negligible.

2. Since the maximum value of α is given by

$$\alpha_{max} = 1 + \frac{R^2}{4Z_o^2} \quad , \qquad (8)$$

Z_o may be calculated in terms of the measured α_{max}.

3. The low frequency capacitance $C = \frac{v}{\ell Z_o}$ can be measured. Since

$$\tau = \frac{\ell}{v} , \quad Z_o = \frac{\tau}{c} \quad . \qquad (9)$$

The measurement techniques are described in the next section.

3.0 EXPERIMENTAL WORK

Insertion loss measurements formed the major part of the experimental work. The very low characteristic impedance of the "transmission line" made it difficult to use the more conventional line techniques, e.g., time-domain reflectometry, etc. The equipment used included a GR-1001A signal generator, a Spencer-Kennedy impulse generator, an NF-105 noise and field intensity meter, several 50-ohm line attenuators, and a Tektronix 569 dual beam oscilloscope.

3.1 TEST PROCEDURE

All the capacitors were fitted with BNC jacks, arranged for minimal lead inductance. The insertion loss measurements were made in the conventional manner (see Fig. 5) [14]. A reading was established on the noise meter with the filter in place. The filter

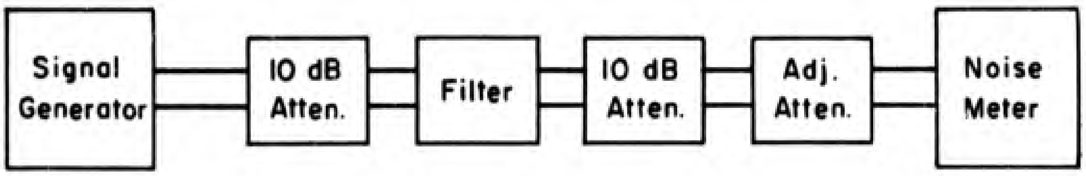

Figure 5 INSERTION LOSS MEASUREMENTS

was removed, and the adjustable attenuator set to restore the original reading on the noise meter. The increase in attenuation required was the insertion loss. The noise meter served only as a level indicator, the calibrated attenuator being the standard of measurement. Initial tests were made using the impulse generator as a signal source, which allows tuning the noise meter to locate any pronounced peaks or dips quickly. The impulse generator is satisfactory as a signal source for regular measurements, except that its limited output may not be sufficient for filters with high attenuation, and system resolution depends on the noise meter's bandwidth.

3.2 SINGLE-ENDED INSERTION LOSS

A modified insertion loss test was used to investigate the input impedances of the filters. Here one end of the filter was attached to a T fitting in the line, which put the filter's input impedance across the line (see Fig. 6). The other port of the filter could be either shorted or left open. This technique

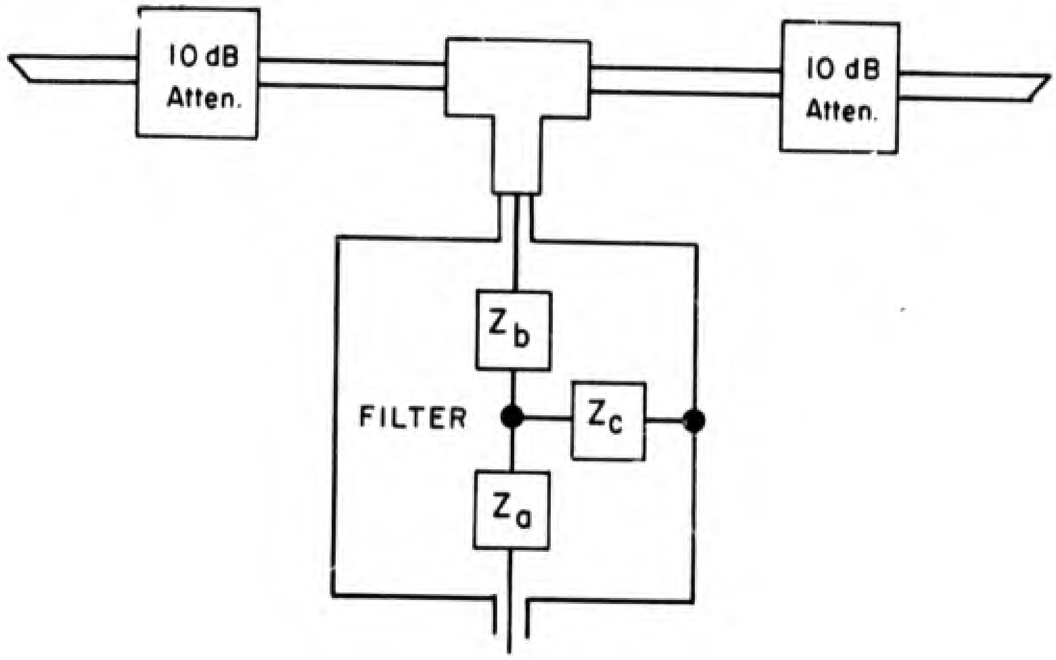

Figure 6 SINGLE-ENDED INSERTION LOSS MEASUREMENT

was also used for tests of ordinary rolled foil capacitors, several of which were tested.

3.3 DELAY MEASUREMENTS

For the feedthrough units, a rather crude measurement of the delay, and hence the electrical length of the line, was made by means of a dual beam oscilloscope. This was done by putting the input signal on one beam of the scope, and the output on the other beam, and observing the phase relation between the two (see Fig. 7). Here again the impedance match is poor so that the capacitor attenuates the signal into the noise, especially at the higher frequencies. But several phase reversals were discernable in most cases.

3.4 TEST RESULTS

The results for a number of capacitors are presented in Figs. 8 - 15.

The curve of Fig. 8 shows the insertion loss of one particular 10 µF feedthrough capacitor. The figure also shows the expected insertion loss of an ideal 10 µF unit, and the performance required by SAE-ARP-936. On the right side is indicated a scale of $|Z|$, the idealized shunt impedance magnitude, corresponding to the scale of insertion loss α. The α curve shows peaks at 0.7 MHz and 2.2 MHz, and has dips at 1.5 MHz and 2.8 MHz. The two clear peaks of α are 60 dB and 57 dB. Figure 9 shows one of the single-ended insertion loss curves for the filter. The other end shows a similar

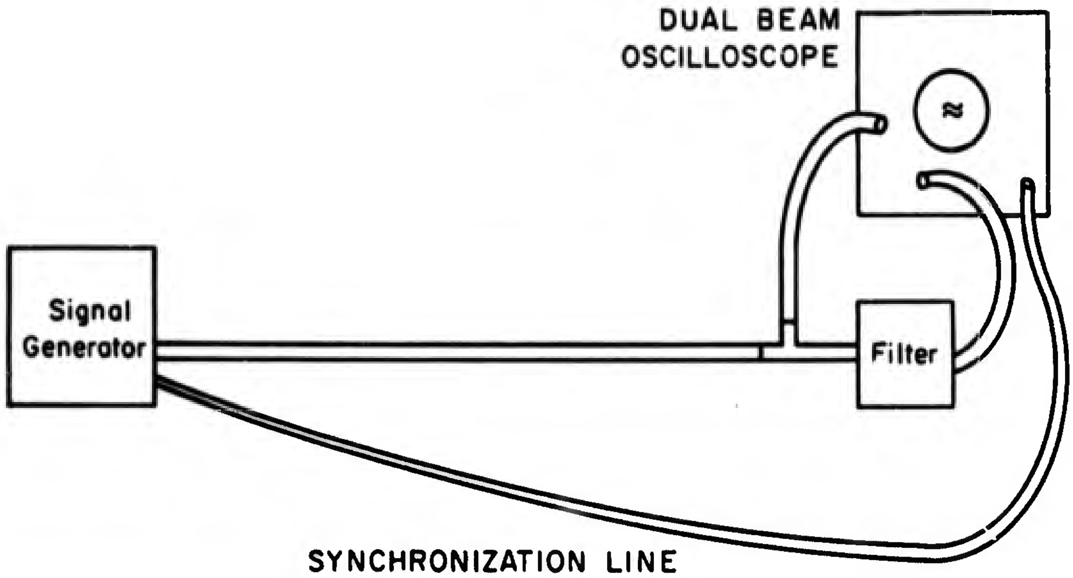

Figure 7 DELAY MEASUREMENT

pattern. They shows peaks characteristic of a simple series resonance in Z. Above the resonant frequency, the insertion loss decreases as $1/f^2$, as would be expected from an inductance across the line (eq. 4). But at 1.5 MHz and 2.8 MHz are seen what appear to be low-Q parallel resonances in the input impedance.

The open and short-circuit input impedances were measured at 50 kHz, yielding a value for Z_o of 0.08 ohms.

The delay of the line was measured and yielded a delay of 0.5 μsec, hence $Z_o = \frac{\tau}{C} = \frac{4 \times 10^{-7}}{10^{-5}} = .04$ ohms. Taking the peak insertion loss as 60 dB or 10^6, eq. 8 yields .025 ohms as the characteristic impedance.

The various results for the characteristic impedance Z_o are:

	estimated	$\sqrt{Z_{oc}Z_{sc}}$	τ/C	from α_{max}
Z_o	0.1 ohm	.08 ohm	.04 ohm	.025 ohm

In view of the uncertainties of the parameters of the transmission line model of the capacitor, the factor of 4 difference between the values of Z_o does not seem excessive.

The effects of lead inductance, discussed in section 2.3, obscure the higher order transmission line resonances, but the dips and peaks are qualitatively accounted for by this model.

Figure 10 shows the measured insertion loss of another manufacturer's 10 μF feedthrough capacitor. The primary dip goes some 40 dB below the expected value, well below even the SAE-ARP-936 curve.

Figure 11 shows the results for similar tests on a 2 μF feedthrough capacitor, manufactured by a third company.

Only one dip in α is discernable in this case, and the associated effect on the input impedance is scarcely visible. This unit was unsymmetrical, as seen by the different series resonance frequencies on the single-ended curves.

Figure 12 shows the results for a 0.25 μF feedthrough unit purported to be effective at VHF. Tests to 100 MHz indicate performance falling far short of ideal, but having only a slight dip at about 50 MHz.

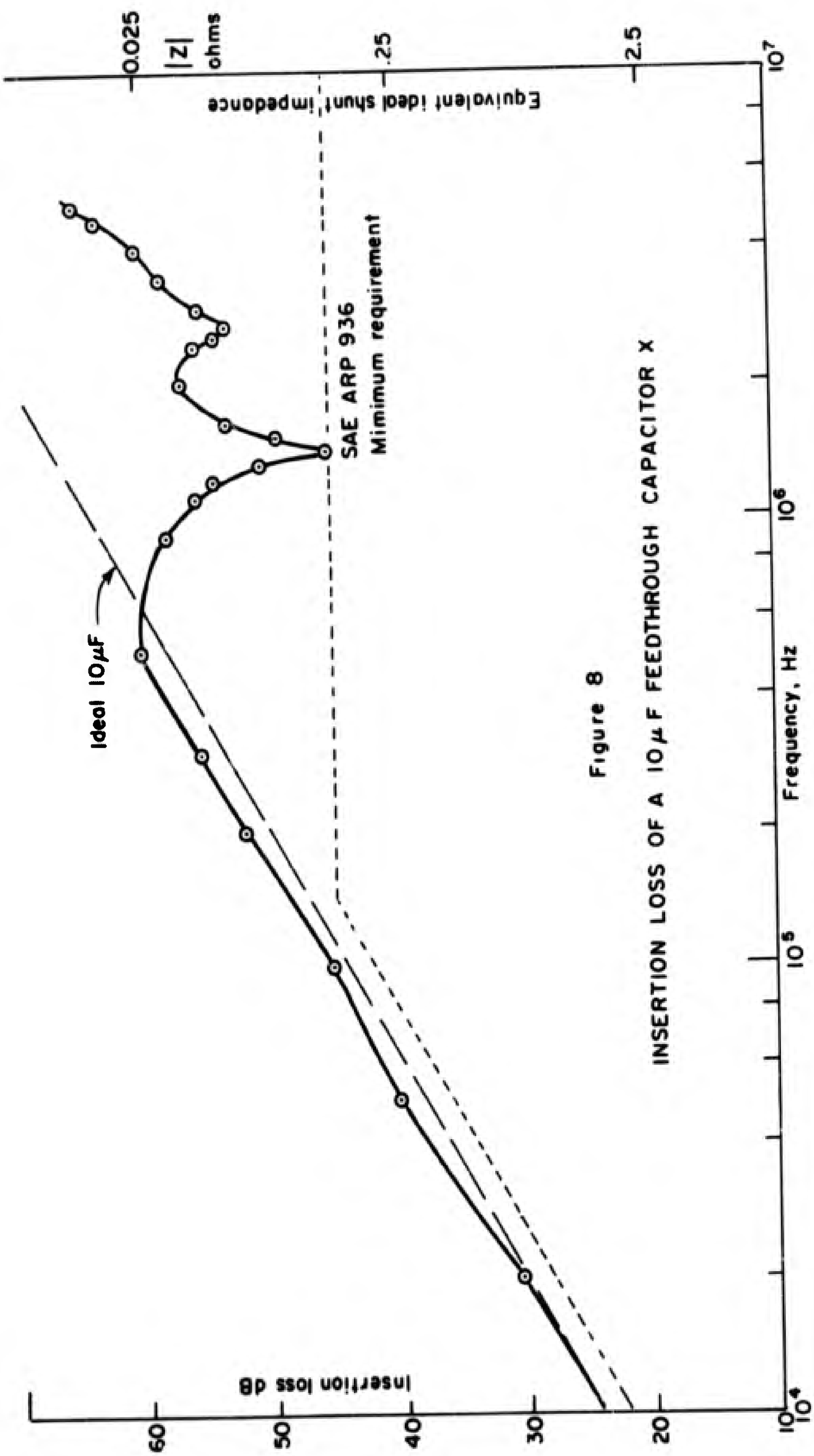

Figure 8

INSERTION LOSS OF A $10\mu F$ FEEDTHROUGH CAPACITOR X

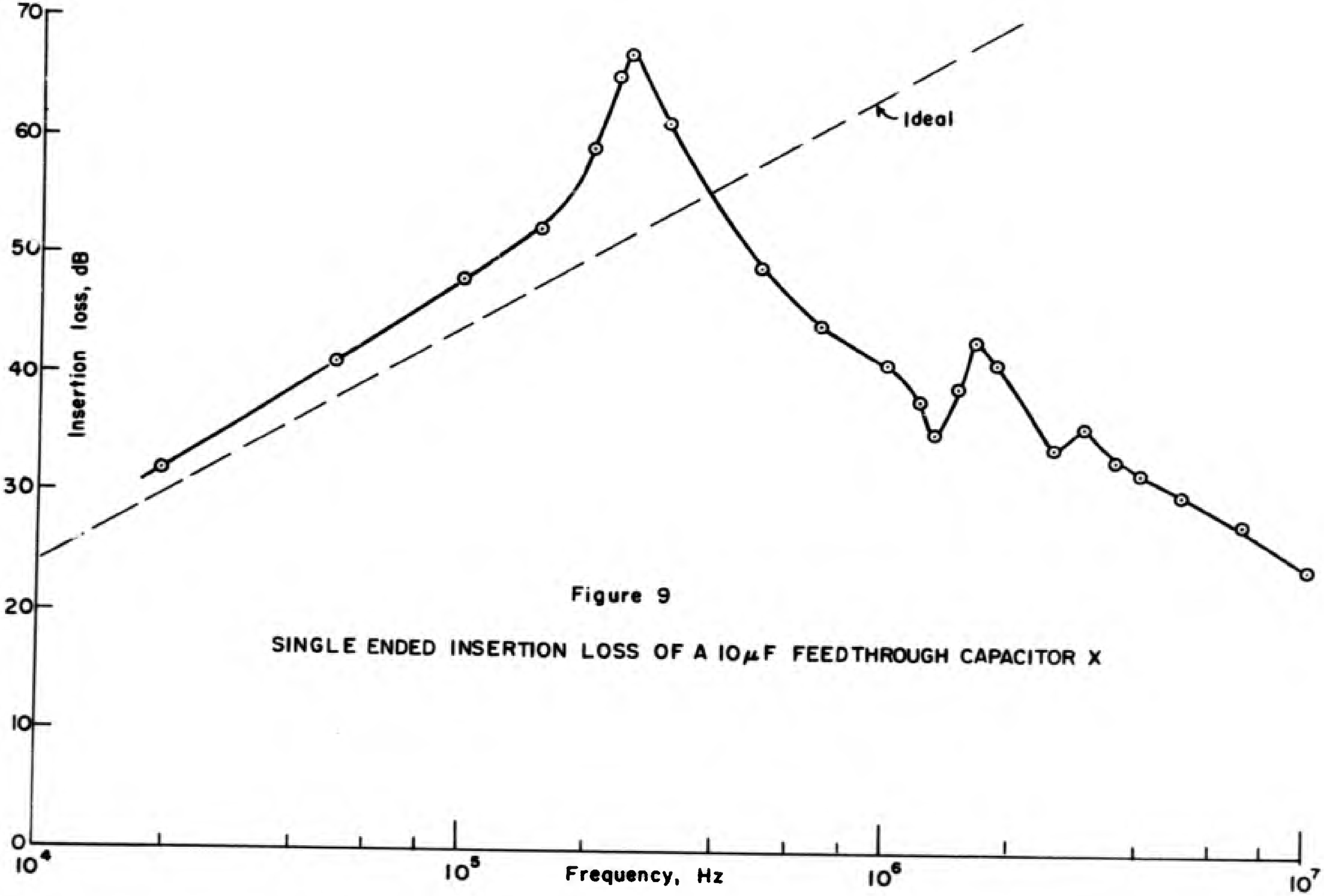

Figure 9

SINGLE ENDED INSERTION LOSS OF A 10μF FEEDTHROUGH CAPACITOR X

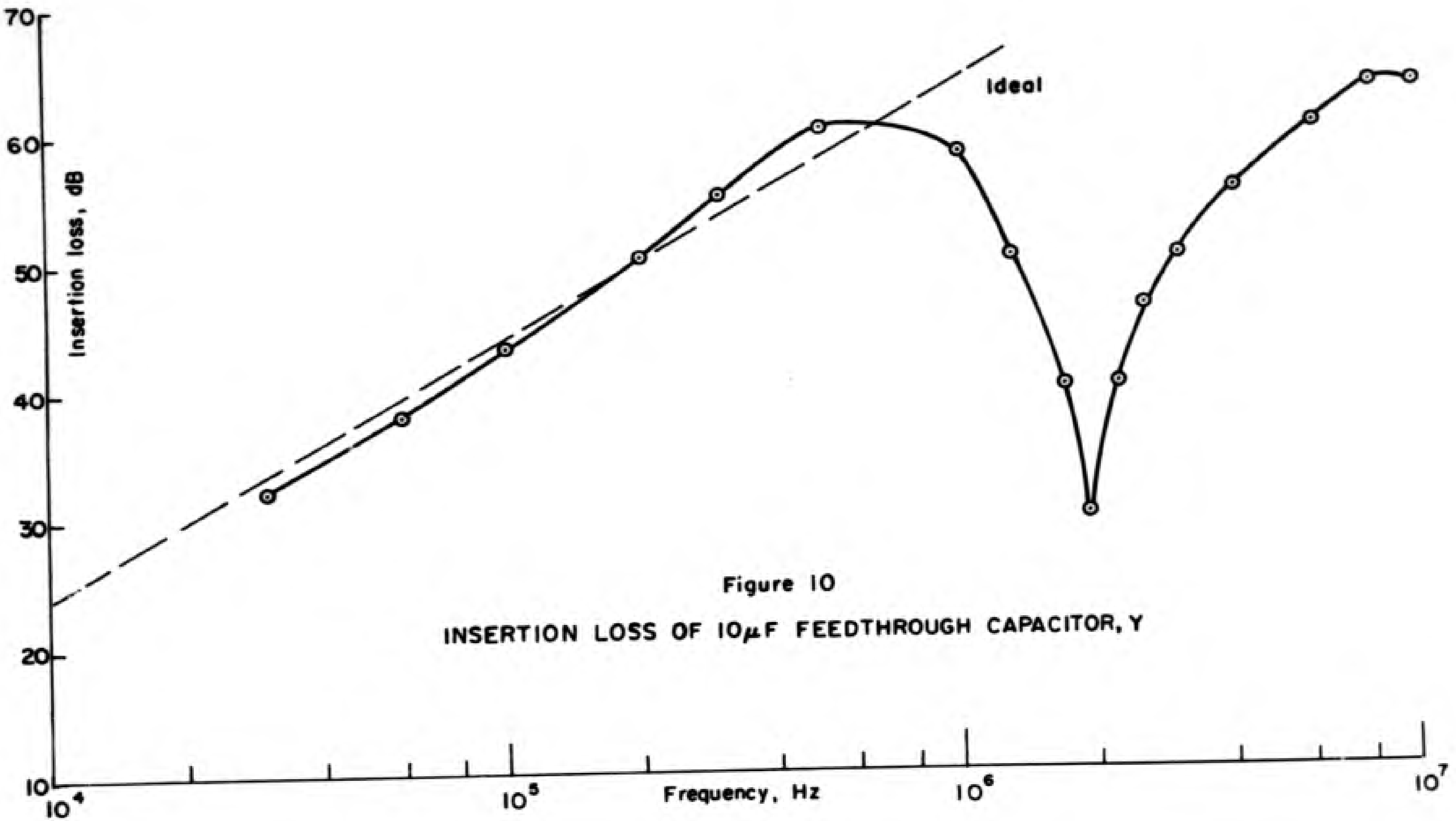

Figure 10

INSERTION LOSS OF 10μF FEEDTHROUGH CAPACITOR, Y

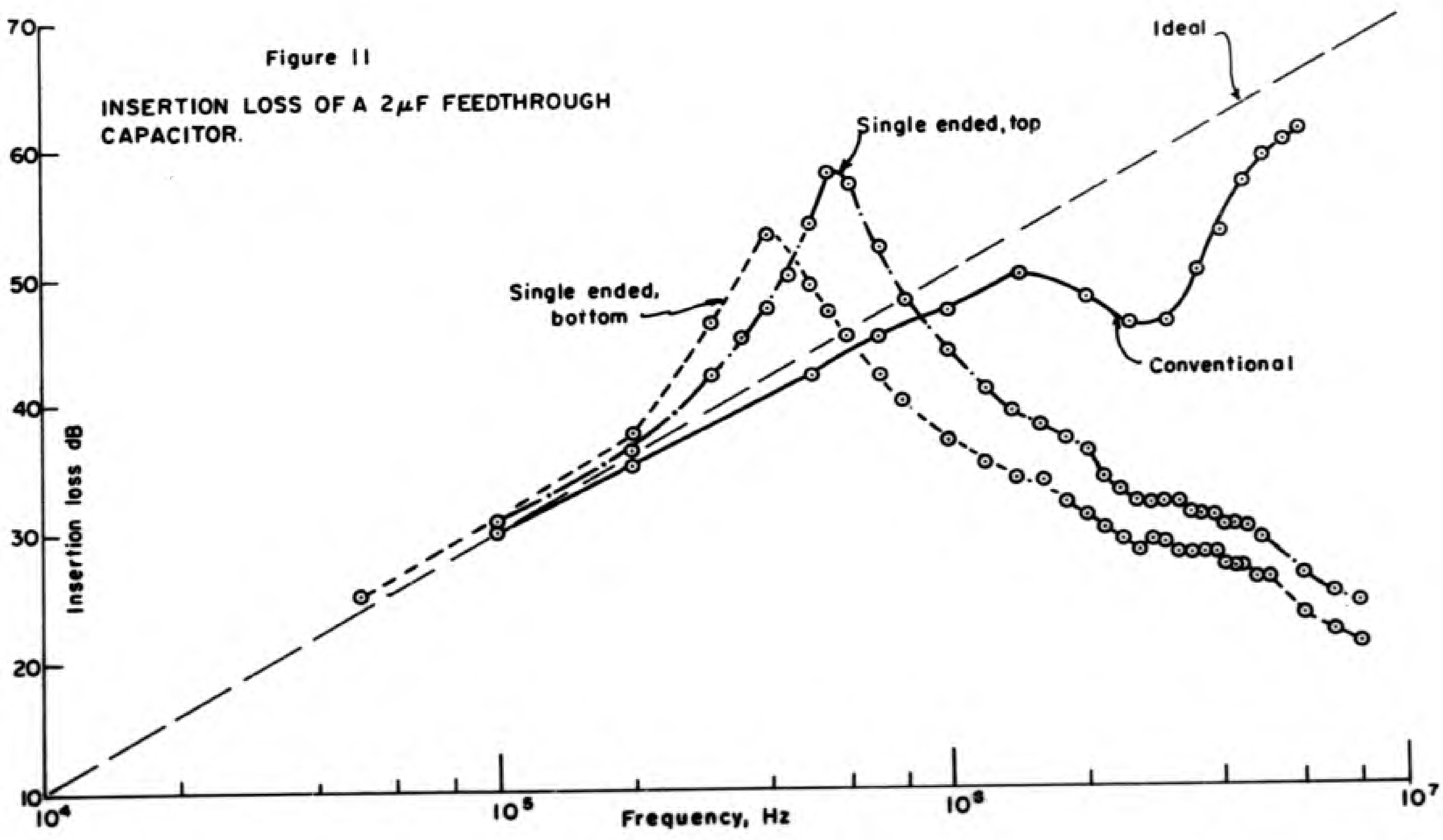

Figure 11

INSERTION LOSS OF A 2μF FEEDTHROUGH CAPACITOR.

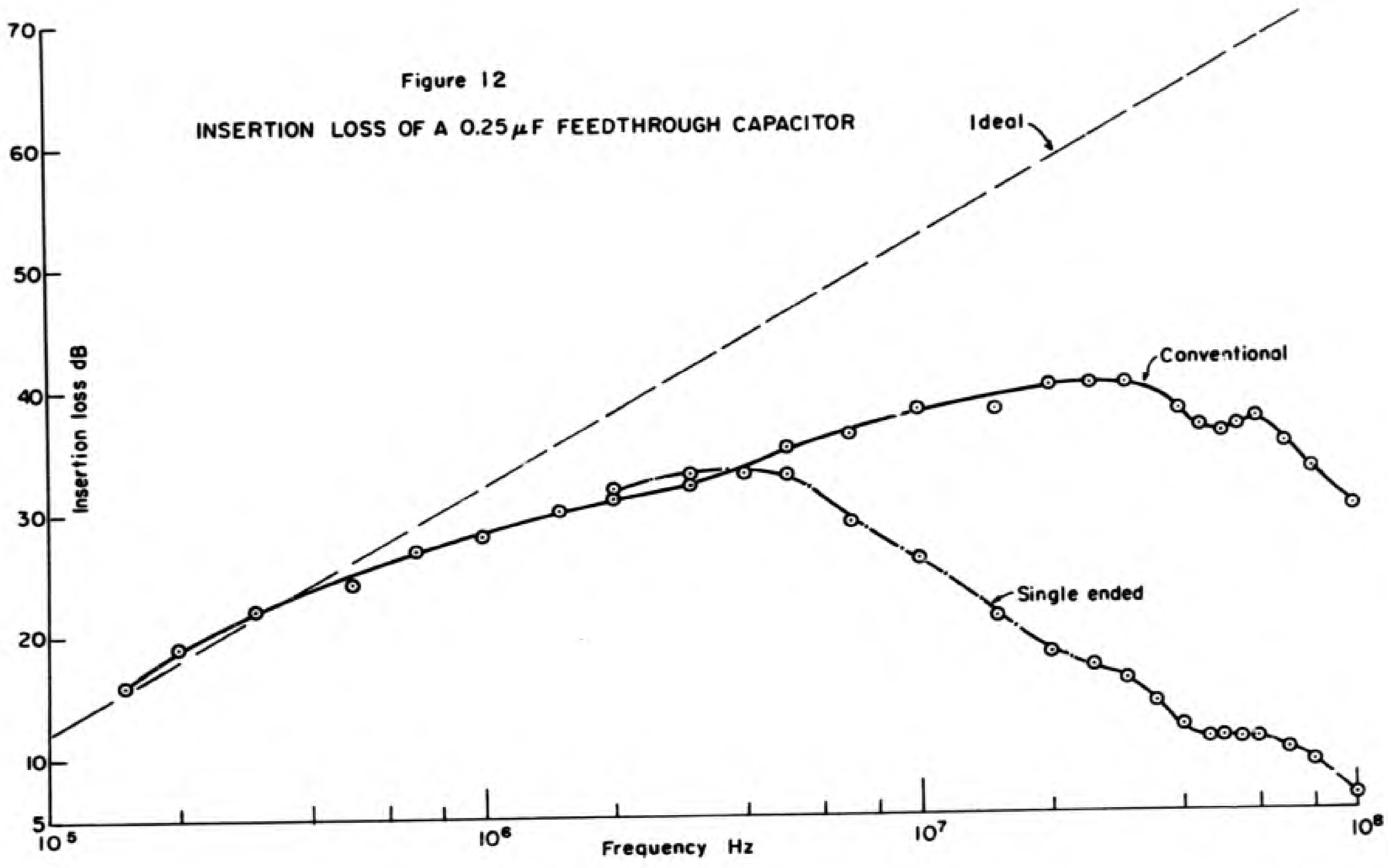

Figure 12

INSERTION LOSS OF A 0.25 μF FEEDTHROUGH CAPACITOR

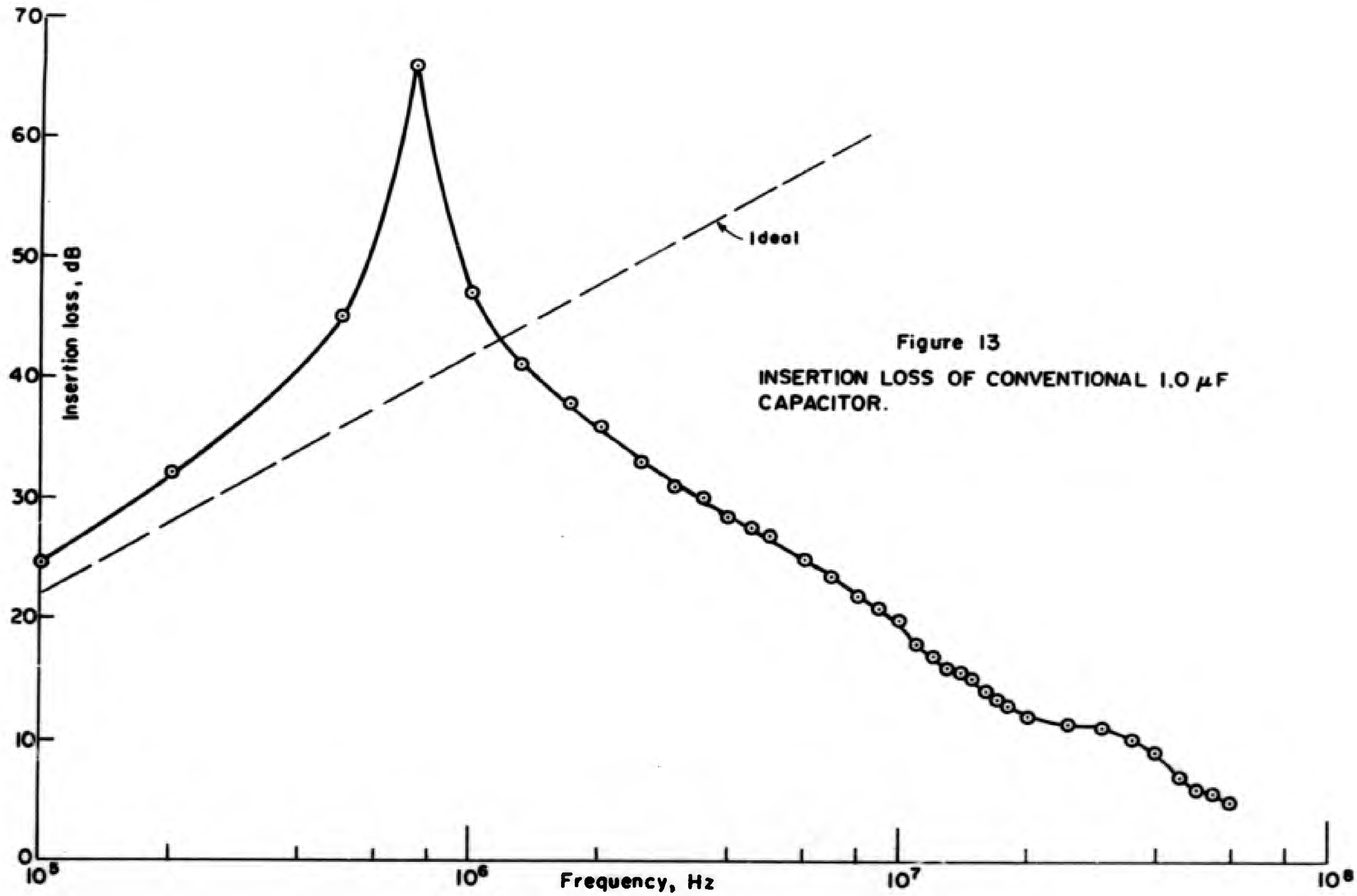

Figure 13

INSERTION LOSS OF CONVENTIONAL 1.0 μF CAPACITOR.

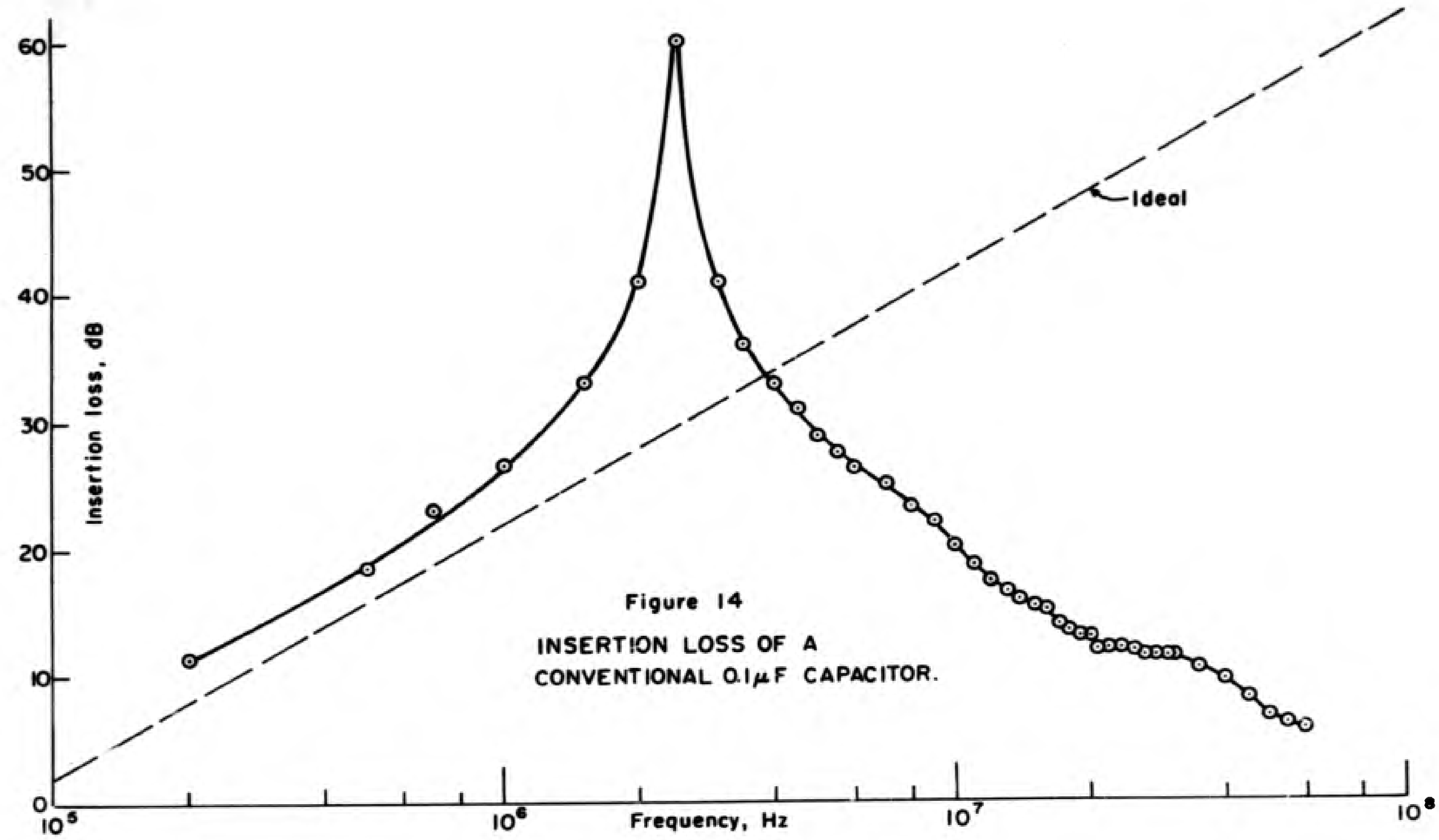

Figure 14

INSERTION LOSS OF A
CONVENTIONAL 0.1 μF CAPACITOR.

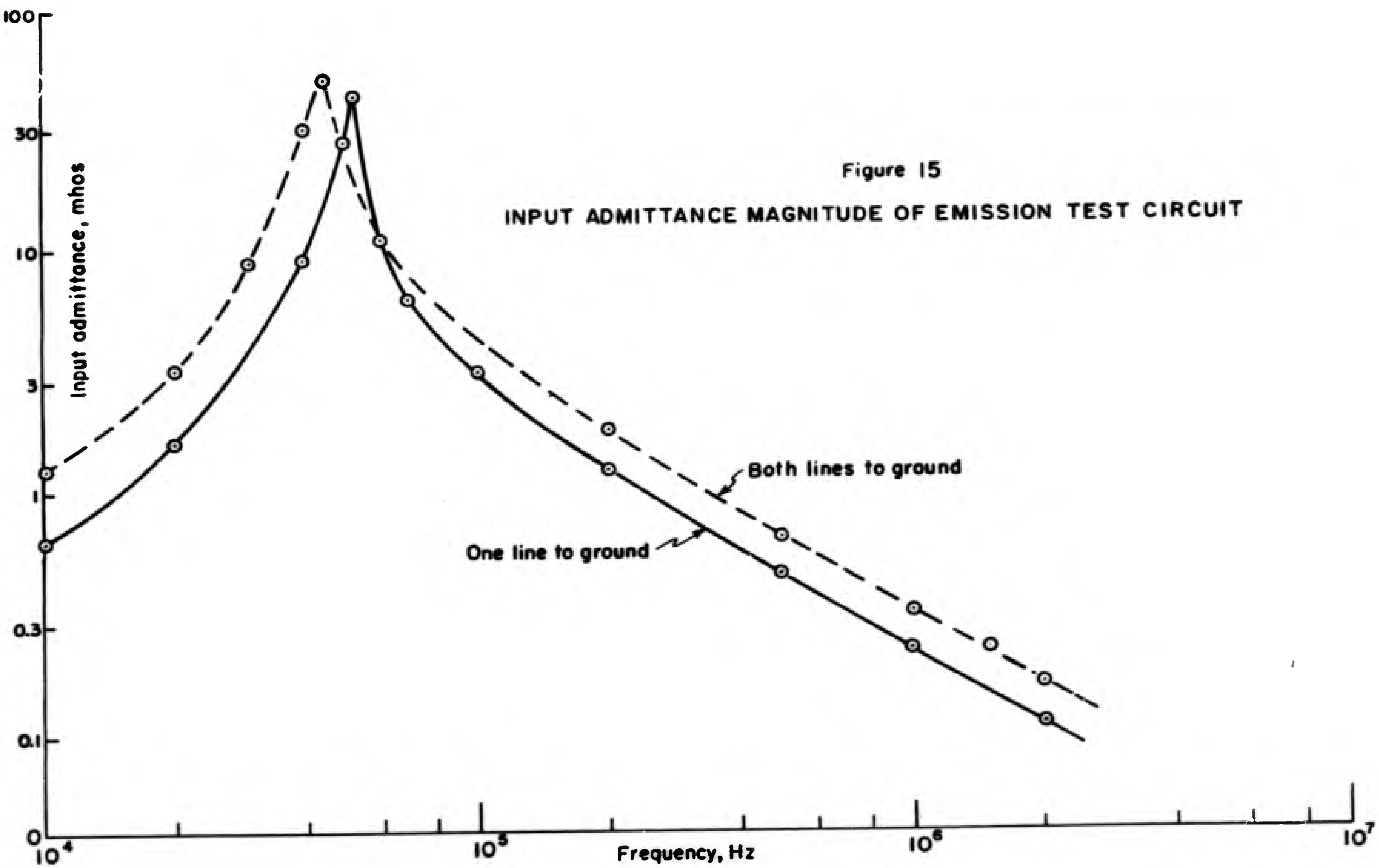

Figure 15

INPUT ADMITTANCE MAGNITUDE OF EMISSION TEST CIRCUIT

This phenomenon has also been detected in conventional rolled foil capacitors. Figures 13 and 14 show the measured insertion loss of a 1.0 μF and a 0.1 μF unit. These units have a mylar dielectric and a plastic case. The series resonance is the obvious feature of the curve, but on the high frequency inductive tail of the curves, we see deviations characteristic of the transmission line resonances. It is difficult to imagine a situation where this might have any practical consequences.

3.5 AN EMISSION TEST CIRCUIT

Figure 16 shows the layout of a circuit for measuring emission currents, using feedthrough capacitors [15]. The equivalent circuit of the arrangement is shown. The input admittances

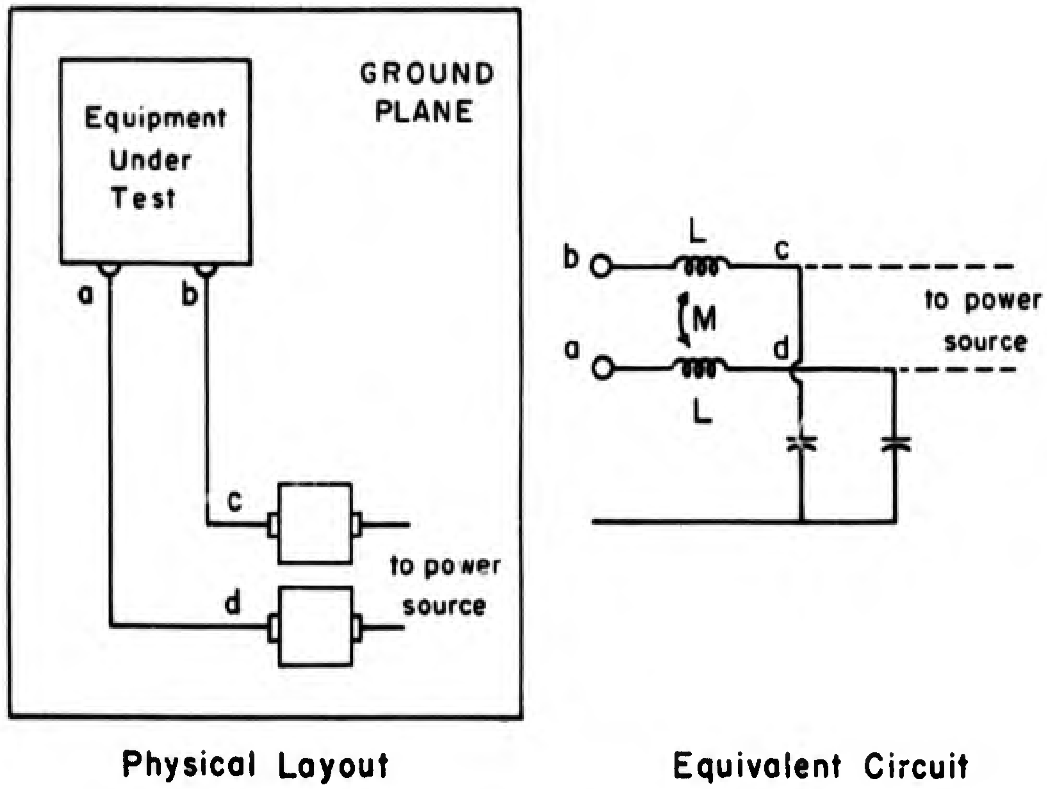

FIGURE 16 TEST CIRCUIT FOR MEASURING EMISSION CURRENTS

presented to the test sample by this circuit were measured, using the single ended insertion loss technique described above. Measurements were made at each terminal, and also with both terminals shorted together. The two inductances are not identical, but the difference is not significant. The curves of Fig. 15 show the magnitude of the input impedance on a logarithmic scale. It can be seen that series resonance occurs at about 50 kHz. Above this, the circuit is primarily the lead inductance, so the capacitor resonances have no significant effect where they occur. The value of L can be determined from the series resonance points; it is about 0.9 μH. The value of the mutual inductance M between the two wires is about 0.3 μH. In the HF region, these quantities are really the important parameters of the setup. At 10 MHz, for example, the impedance between a and b is approaching 100 ohms. It is quite possible that this might be more of an open circuit than a short circuit, when compared to the equipment's output impedance.

4.0 CONCLUSIONS AND RECOMMENDATIONS

It is clear that large feedthrough capacitors show transmission line characteristics in the HF range. The complexity of the transmission line involved has been indicated. The heart of the matter is that the shunt reactance of the capacitor is better described by $\dfrac{1}{j\dfrac{\sin \omega\tau}{\tau}C}$ than by $1/j\omega C$, where τ is the delay time of the capacitor.

The isolation of the test sample from the power source provided by the capacitors falls short of what would be expected from 10 μF. It should be made clear that additional isolation is required if spurious signals from the power source are present. Typically, this would be provided by the line filters of a shielded enclosure, but these may not always be present.

As an RF short for the emission test setup for CE01 and CE03 of MIL-STD-462, these capacitors are as satisfactory as an ideal 10 μF element. But it should be understood that the inductive reactance of the separated power leads will dominate the impedance above about 50 kHz, which is the series-resonant frequency.

In the MF and HF regions, series resonance is possible at higher frequencies if the output impedance of the test sample is capacitive. Use of the line impedance stabilization network (LISN) has the advantage of damping such resonances and yielding more representative measured data.

It is specifically recommended that:

1. An auxiliary requirement on insertion loss be imposed to assure that ambient conducted interference will not affect the measurement. Particular attention should be given to values of insertion loss at frequencies below 10 kHz and above 1 MHz. (SAE-ARP-936 is included in proposed revision to MIL-STD-462. This standard provides for minimum requirements on insertion loss as a function of frequency. The requirement given there may or may not be adequate in any given setup.)

2. Further investigation of the dependence of the measured currents at frequencies above 1 MHz on the particular 10 μF capacitor used should be made, with a view toward substituting a network which contains dissipative elements in order to avoid resonance effects.

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Appendix I

TRANSMISSION LINE MODEL

Consider a length ℓ of a transmission line of low impedance Z_o, assumed to be uniform and lossless. The input impedance [8] of an open-circuited length of line is $-j\, Z_o \cot \beta\ell$ where $\beta = \frac{\omega}{v}$, and v is the velocity of wave propagation on the line. The small angle approximation for cot X is just $\frac{1}{X}$, so the low frequency limit for Z is $-j\, Z_o \frac{v}{\omega\ell}$. So the low frequency capacitance is equivalent to $C = \frac{\ell}{vZ_o}$.

For a parallel strip line of width W and separation d ($d \ll W$), the characteristic impedance [9] is $\sqrt{\frac{\mu}{\epsilon}}\,\frac{d}{W}$. Since $v = \frac{1}{\mu\epsilon}$, we have that $C = \epsilon\,\frac{\ell W}{d}$, a familiar result. But for higher frequencies, the capacitor would have an input impedance with the characteristic set of poles and zeros, and would exhibit a delay τ equal to ℓ/v.

The computation of the insertion loss of this filter is facilitated by the T section equivalents for a line section [10]. They are:

$$Z_a = Z_b = j\, Z_o \tan\frac{\beta\ell}{2} = j\, Z_o \frac{\sin\beta\ell}{1+\cos\beta\ell}$$

$$Z_c = -j\, Z_o \cos\beta\ell = -j\, Z_o \frac{1}{\sin\beta\ell}$$

The expression for α, eq. (1), with $Z_a = Z_b$, can be written

$$\alpha = \left|1 + \frac{Z_a}{Z_c} + \frac{Z_a^{\,2} + 2Z_aZ_c}{2RZ_c} + \frac{R}{2Z_c}\right|^2$$

Substituting the above expressions, we have

$$\alpha = \left| 1 - \frac{\sin^2 \beta\ell}{1 + \cos \beta\ell} + \frac{-Z_o^{\,2} \dfrac{\sin^2 \beta\ell}{(1 + \cos \beta\ell)^2} + 2Z_o^{\,2} \dfrac{1}{1 + \cos \beta\ell}}{-2RjZ_o \dfrac{1}{\sin \beta\ell}} + j \frac{R \sin \beta\ell}{2Z_o} \right|^2$$

which reduces to

$$\alpha = \left| \cos \beta\ell + j \left(\frac{R}{2Z_o} + \frac{Z_o}{2R} \right) \sin \beta\ell \right|^2$$

and this is equal to

$$\alpha = \cos^2 \beta\ell + \left(\frac{R}{2Z_o} + \frac{Z_o}{2R} \right)^2 \sin^2 \beta\ell \ .$$

Since $\cos^2 \alpha = 1 - \sin^2 \alpha$,

$$\alpha = 1 + \left(\frac{R}{2Z_o} - \frac{Z_o}{2R} \right)^2 \sin^2 \beta\ell \ .$$

Now if $Z_o \ll R$, as is likely

$$\alpha = 1 + \frac{1}{4} \frac{R^2}{Z_o^{\,2}} \sin^2 \beta\ell = 1 + \frac{1}{4} \frac{R^2}{Z_o^{\,2}} \sin^2 \omega \frac{\ell}{v} \quad (6)^*$$

We expect the insertion loss to drop to unity at $\omega = n\pi \frac{\ell}{v} = n \frac{\pi}{\tau}$. At these same frequencies we expect a pole in the input impedance. The insertion loss should have a maximum value of $1 + \frac{R^2}{4Z_o^{\,2}}$ at $\omega = (n + \frac{1}{2}) \frac{\pi}{\tau}$.

---

* Equation numbers in this appendix correspond with those used in the text.

A.1 VALIDITY OF THE TRANSMISSION LINE MODEL

How closely does a rolled foil capacitor approximate a length of transmission line? The lengthwise cross-section of an extended foil unit is essentially this:

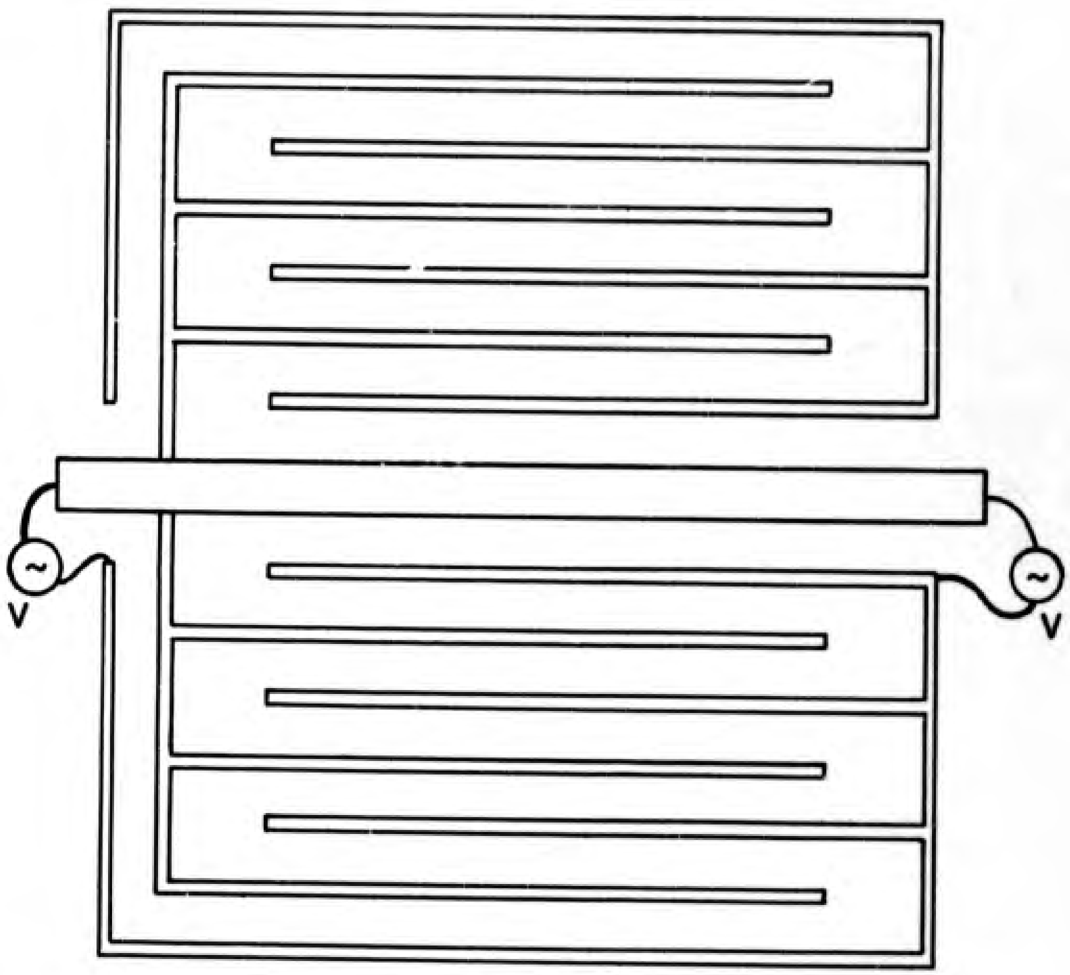

Figure A.1 SIMPLIFIED CROSS-SECTION OF ROLLED FOIL FEEDTHROUGH CAPACITOR

An excitation applied at the left end of the structure will cause an electromagnetic wave to pass back and forth between the foils, moving parallel to the axis, and getting closer to the axis. (If started from the right, the wave progresses outward, see Fig. A.1.) From this point of view, the rolled structure is a coaxial line folded into itself many times.

There are three major difficulties in making this analysis more precise:

1) The line is nonuniform.

2) Segments of the line are mutually coupled.

3) Apparently $j\omega L \ll R$ per unit length.

A.1.1 NONUNIFORMITY OF LINE

This line is nonuniform in two respects. If the circumference at any point is thought of as the width of a parallel strip line, the line is narrowest at the inside, and has an impedance which decreases with increasing radius. (The characteristic impedance of a parallel strip line of width W and separation d is $\sqrt{\frac{\mu}{\epsilon}}\,\frac{d}{W}$.) Also at the end of each segment, the overlapping region adds a lumped inductive component at this point. But these occur so frequently that they may be thought of as part of the distributed inductance of the line.

A.1.2 MUTUAL COUPLING

The simple notion of voltage waves traveling through the structure is valid only if the associated currents flow on the surface of the conductors. But the skin depth [11] in aluminum at 1 MHz is 8×10^{-2} cm, and the foil used in such capacitors is typically 10^{-3} cm thick. So the wave in one segment is not independent of the wave in the next segment, but instead the two are cross-coupled by the mutual currents.

A.1.3 EXCESSIVE LOSS

A representative thickness for the dielectric is 2×10^{-3} cm. If we consider a segment of the line as a parallel strip line, a representative width would be a representative circumference, say $\pi \times 2$ cm. Assuming a dielectric constant of 3 we can compute the following parameters of the line (all per unit length).

$$\ell = \mu \frac{d}{w} \qquad r = \frac{\rho}{tw}$$

$$c = \epsilon \frac{w}{d} \qquad g = \tan\delta\,\omega\,\epsilon\,\frac{w}{d}\,.$$

Here ρ is the resistivity of aluminum, and $\tan \delta$ is the loss tangent of the dielectric, d is the conductor separation, and t is the conductor thickness. For a general transmission line [12].

$$Z_o = \sqrt{\frac{R + j\omega L}{G + j\omega C}}, \qquad \delta = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$$

For this case

$$R + j\omega L = \frac{\rho}{tw} + j\omega\mu \frac{d}{w} = \frac{2.8 \times 10^{-6} \text{ohm cm}}{10^{-3}\text{cm} \times 2\pi \text{ cm}} + j\omega \frac{4\pi\ 10^{-7}\text{h}}{\text{m}} \cdot \frac{2 \times 10^{-3}\text{cm}}{2\pi \text{ cm}}$$

$$= 5 \times 10^{-4} \text{ ohm/cm} + j\omega\ 4 \times 10^{-12} \frac{\text{h}}{\text{m}}$$

at 1 MHz, $\omega = 2\pi\ 10^6 \text{ sec}^{-1}$

$$R + j\omega L = 5 \times 10^{-4} + j\ 2.5 \times 10^{-5} \text{ ohm/cm}$$

The $j\omega L$ term appears small compared to the distributed resistance of the line (even without skin effect).

Considering

$$G + j\omega C = \tan \delta\ \omega\ \epsilon \frac{w}{d} + j\omega\epsilon \frac{w}{d} = \omega\epsilon \frac{w}{d} (\tan \delta + j)$$

it is unlikely that $\tan \delta$ would approach 1, so $j\omega C$ is a good approximation for the shunt term.

It would seem that this is an RC type of line rather than an LC type. The periodic folds provide some additional inductance, but the contribution is of the same order of magnitude as the distributed inductance.

A.2 CHARACTERISTIC IMPEDANCE OF THE LINE

How can one estimate the value of the characteristic impedance of the line? The characteristic impedance of a parallel strip transmission line is $\sqrt{\frac{\mu}{\epsilon}} \frac{d}{w}$. Inserting representative values, we find that the characteristic impedances should be of the order of 0.1 ohm.

From transmission line theory, we know that

$$Z_o = \sqrt{Z_{oc} Z_{sc}} \qquad (7)$$

where Z_{oc} and Z_{sc} are the open-circuit and short-circuit input impedances [13]. These measurements must be made at a low enough frequency for the lead reactance to be negligible. since the actual filter has stray inductances at each end of the capacitive structure.

If clear peaks in the insertion loss are found, correspon-ding to eq. 5, then Z_o may be found from

$$\alpha_{max} = 1 + \frac{R^2}{4Z_o^{\ 2}} \tag{8}$$

In section 2.3 it was shown that the low frequency capacitance is given by $C = \frac{v}{\ell Z_o}$. The delay $\tau = \frac{\ell}{v}$ is an easily measurable quantity. So we have

$$Z_o = \frac{\tau}{C} \tag{9}$$

as a third way to measure Z_o.

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13. ABSTRACT

The insertion loss of a number of power line feedthrough filter capacitors shows pronounced deviation from the expected results in the HF region, which is traced to the low impedance transmission line character of the filter.

The insertion loss of the filter is predicted from models of varying degrees of sophistication. The validity of considering a rolled foil capacitor as a transmission line is discussed, the theory indicating a series of dips in the insertion loss, and a series of poles and zeroes in the input impedance.

Experimental results clearly indicate transmission line effects in a number of feedthrough units. Similar behavior was also detected in conventional rolled foil capacitors. The practical significance of the results is discussed along with the use of such filters for measuring spurious emission from electronic equipment.

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