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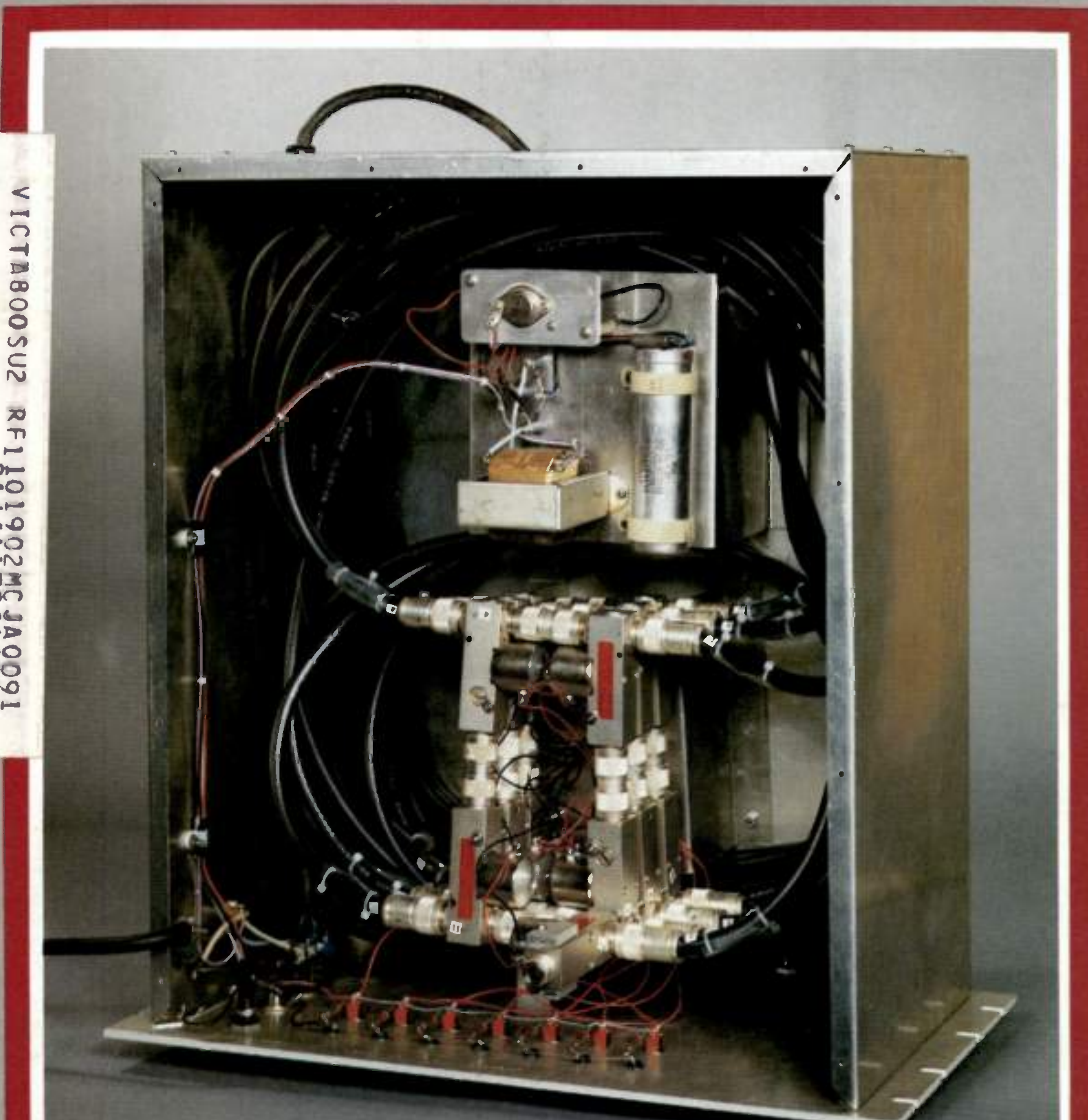

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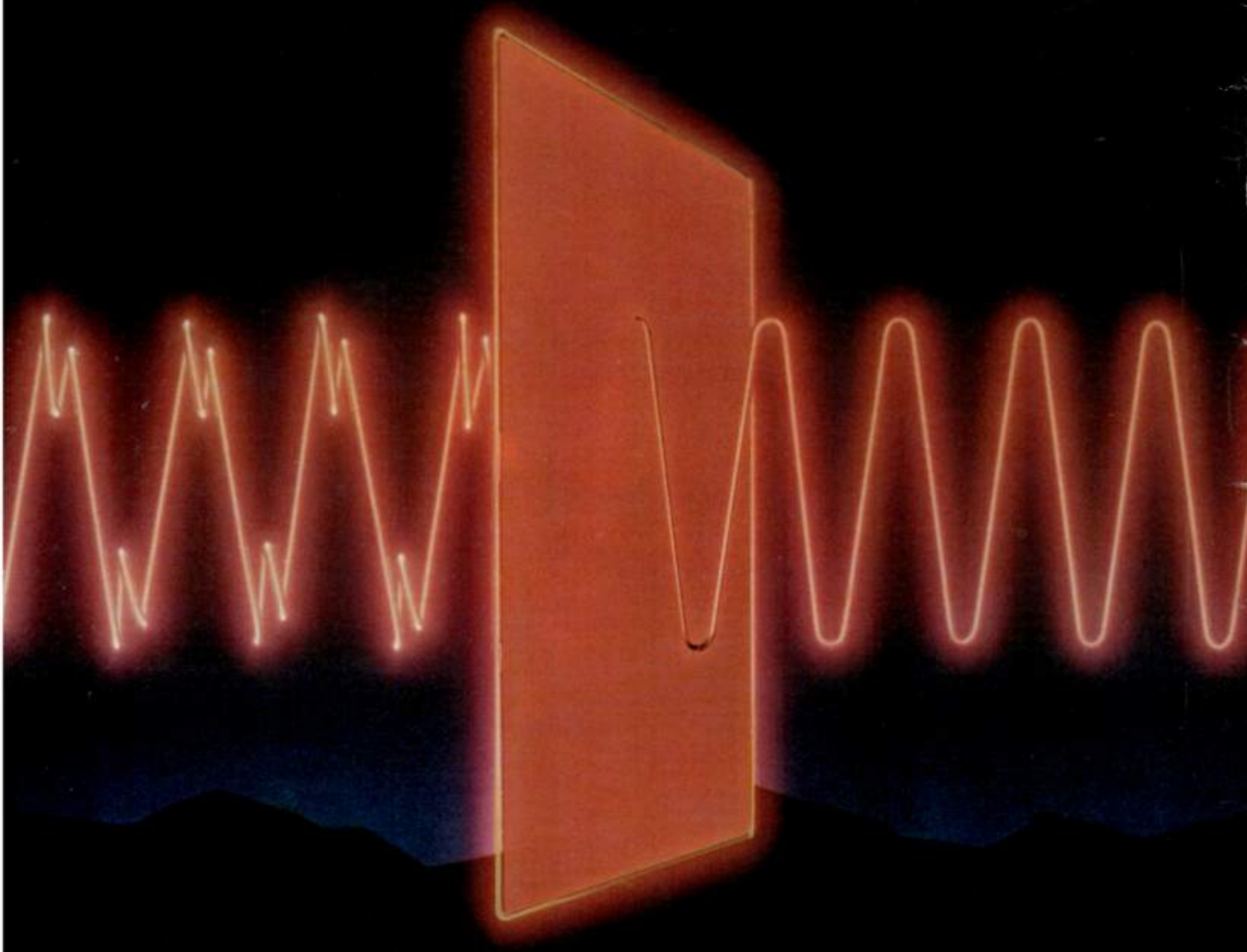

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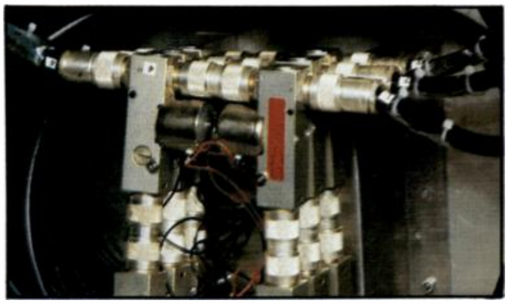

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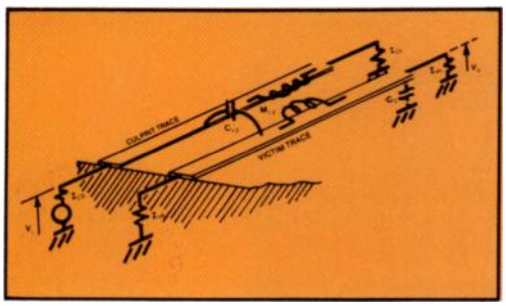

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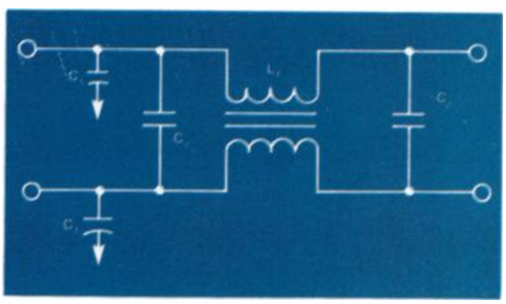

Insertion Loss Testing

July/August 1982, Volume 5, No. 4, **r.f. design** (ISSN 0163-321X) is published bi-monthly by Cardiff Publishing Company, a subsidiary of Cardiff Communications, Inc., 6430 S. Yosemite St., Englewood, Colo. 80111. (303) 694-1522. Copyright © 1982 Cardiff Publishing Company. Second Class postage paid at Englewood, Colorado, and additional offices. Contents may not be reproduced in any form without written permission. Please address subscription correspondence and Postmaster, please send PS form 3579 to P.O. Box 1077, Skokie, IL 60777. Subscriptions: Canada & Mexico; foreign $20 per year. Please make payment in U.S. funds only. Single copies available at $3 each.

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—AT—6	6	±0.3dB	DC-1500	0.6dB	0.8dB	1.3:1	1.5:1
—AT—10	10	±0.3dB	DC-1500	0.6dB	0.8dB	1.3:1	1.5:1
—AT—20	20	±0.3dB	DC-1500	0.6dB	0.8dB	1.3:1	1.5:1

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EMC/EMI

This issue is focused on a subject which is on the minds of much of the electronics industry at the present time: EMC/EMI. There seems to be concern over the requirements, new regulations, and the best methods for compliance. We attempt to answer a few of these questions in this issue.

Also in this issue, we are kicking off a contest to help our advertisers determine which advertisements are best remembered. For you, it is a chance to end up with a $1,000 savings bond. So look for more details on page 42.

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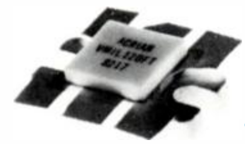

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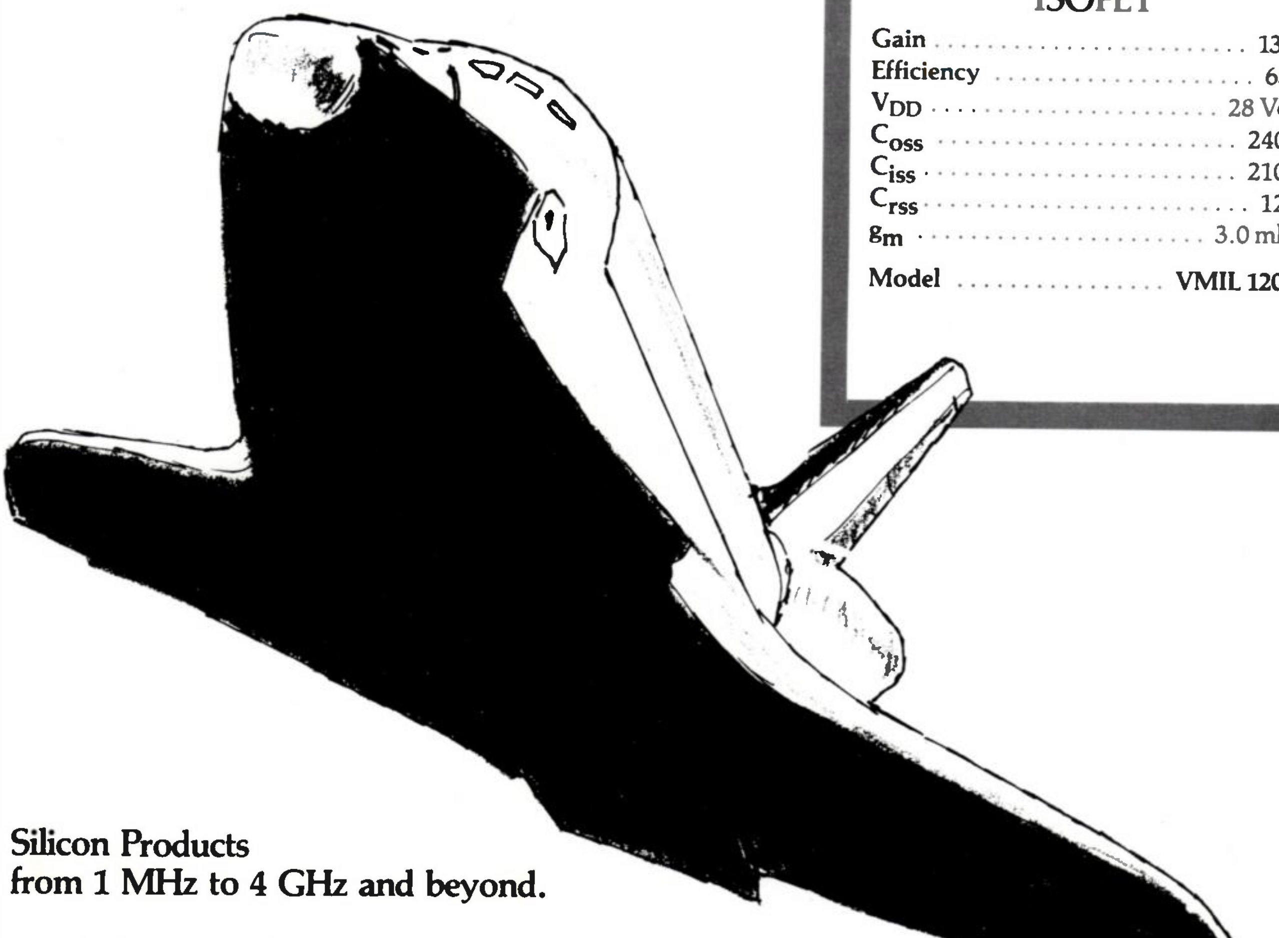

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Prediction and Control of Cross

Crosstalk: A Somewhat Misused Term.

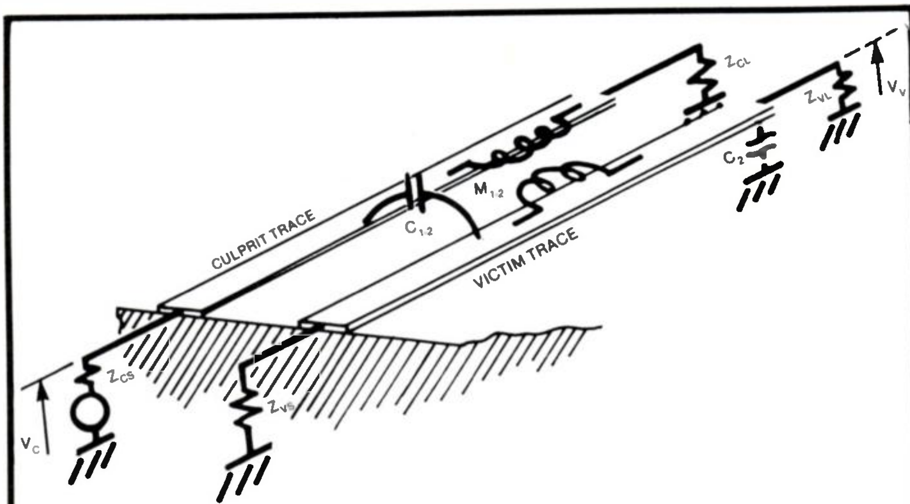


Figure 1. Simplified Crosstalk Equivalent Circuit

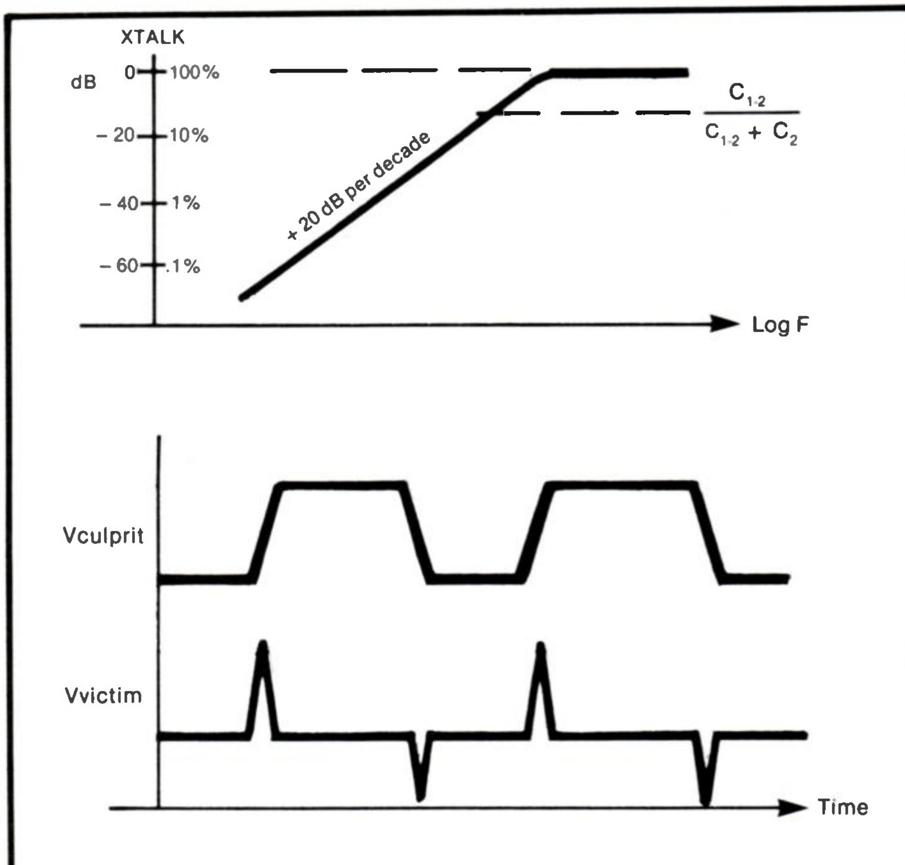


Figure 2. Frequency and Time Domain Illustration of Crosstalk.

By Michel Mardiguian
Director of Training
Don White Consultants, Inc.
Gainesville, VA 22065

Inherited from telephony, where it addresses the parasitic coupling from one voice channel to an adjacent one, the term *CROSSTALK* is employed to characterize, more generally, the percentage of a signal which is coupled from a circuit #1 to a nearby circuit #2 by wire-to-wire influence. Sometimes its meaning has degenerated and, outside the EMC community, some authors use it to cover any kind of undesired noise generation within a limited area, whether it is real wire-to-wire coupling, common impedance coupling or mismatch reflections.

In this article, we will discuss the crosstalk in printed circuit boards strictly in the significance of this term, i.e., the capacitive or inductive transfer between adjacent printed traces.

A Simple Model Permits Quick but Satisfactory Approximations

At the risk of oversimplifying, the PCB crosstalk problem can be represented by Figure 1. Two circuits that we identify as *culprit* and *victim* are using parallel printed traces of some length. As long as the trace is electrically short, i.e., the trace length is << $\lambda/4$ of the culprit frequency or, more precisely, the *rise time* τ_r *of the culprit signal* is longer than the line propagation delay τ_d, the distributed trace-to-trace capacitance and mutual inductance can be replaced by a localized capacitance $C_{1\text{-}2}$ and a localized mutual inductance $M_{1\text{-}2}$.

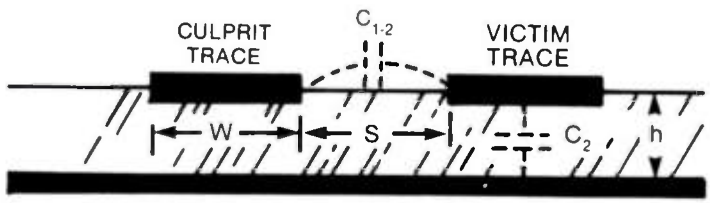
Then it appears that the parasitic voltage V_v transferred to the victim by culprit line voltage V_c has *two contributors:*

(1) The capacitive coupling

(2) The mutual inductance coupling

If we define the crosstalk as being the ratio of $V_{victim}/V_{culprit}$, we have a *figure of merit of any given configuration*, i.e., *how much voltage appears on victim load per volt on the culprit line.* Since telephony and the EMC community uses the decibel extensively, the Xtalk will then be:

talk in Printed Circuit Boards



W/S (Ccv, pF/cm)	W/h = 10 (Cvg = 3.5pF/cm)			W/h = 3 (Cvg = 1pF/cm)			W/h = 1 (Cvg = .35pF/cm)			W/h ≤ .3 or no gnd return below traces			
	.3 (.10)	1 (.20)	3 (.27)	.3 (.10)	1 (.20)	3 (.27)	.3 (.10)	1 (.20)	3 (.27)	.3 (.10)	1 (.20)	3 (.27)	
FREQ. 1 kHz	−144	−138	−136	−144	−138	−136	−144	−138	−136	−144	−138	−136	Culprit
3kHz	−134	−128	−126	−134	−128	−126	−134	−128	−126	−134	−128	−126	Pulse
10kHz	−124	−118	−116	−124	−118	−116	−124	−118	−116	−124	−118	−116	Rise Time
30kHz	−114	−108	−106	−114	−108	−106	−114	−108	−106	−114	−108	−106	10μs
100kHz	−104	−98	−96	−104	−98	−96	−104	−98	−96	−104	−98	−96	3μs
300kHz	−94	−88	−86	−94	−88	−86	−94	−88	−86	−94	−88	−86	1μs
1MHz	−84	−78	−76	−84	−78	−76	−84	−78	−76	−84	−78	−76	300ns
3MHz	−74	−68	−66	−74	−68	−66	−74	−68	−66	−74	−68	−66	100ns
10MHz	−64	−58	−56	−64	−58	−56	−64	−58	−56	−64	−58	−56	30ns
30MHz	−55	−49	−47	−54	−48	−46	−54	−48	−46	−54	−48	−46	10ns
100MHz	−46	−40	−38	−44	−38	−36	−44	−38	−36	−44	−38	−36	3ns
300MHz	−39	−33	−31	−36	−30	−28	−35	−30	−28	−34	−30	−26	1ns
1GHz	−34	−28	−26	−28	−22	−20	−26	−20	−18	−24	−18	−16	.3ns
3GHz	−32	−26	−24	−23	−17	−16	−15	−12	−11	−14	−8	−6	.1ns
10GHz	−31	−25	−23	−21	−15	−14	−13	−10	−8	−4	0	0	.03ns

NOTES
- Epoxy glass assumed ($\varepsilon_r \cong 4$)
- Xtalk given as $20 \text{ Log}_{10} \frac{V_{victim}}{V_{culprit}}$ per cm of parallel run. For other lengths, add $20 \text{ Log}_{10} \ell_{cm}$
- For Z victim ≠ 100Ω (10 to 300Ω), add $20 \text{ Log}_{10} \frac{Z_{victim}}{100}$
- Clamp to 0dB, Xtalk cannot be positive
- Example of some typical values W = 20mils (.5mm), S = 30 mils (.75 mm), for single layer h = 30 − 40 mils (.7 to 1mm), for Multilayer h = 5 mils (.12mm).

Figure 3. Capacitive XTalk Table for 100Ω Total Victim Impedance.

$$\text{Xtalk} = 20\ \text{Log}_{10}\frac{V_{victim}}{V_{culprit}} \quad (1)$$

For the capacitive contributor, a crude expression of the coupling is:

$$\text{Xtalk (cap)} = 20\ \text{Log}\frac{Z_v}{Z_v + \dfrac{1}{jC_{1\text{-}2}\omega}} \quad (2)$$

Z_V, the victim circuit impedance, includes the trace-to-ground capacitance C_2. At frequencies such as $1/C_2\omega >> R_{victim}$, the Xtalk simplifies as:

$$\text{Xtalk (cap)} = 20\ \text{Log}\ R_{victim} \times C_{1\text{-}2}\omega \quad (3)$$

$$\text{or Xtalk (cap)} = 20\ \text{Log}\ 2\pi f \times R_{vict} \times C_{1\text{-}2}$$

If one prefers time domain to frequency domain, the voltage induced in the victim trace can also be expressed as:

$$V_{victim} = R_{vict} \times C_{1\text{-}2} \times \frac{\Delta V}{\tau_r} \quad (4)$$

where ΔV is the culprit voltage swing and τ_r its rise time – with $C_{1\text{-}2}$ in Farads, τ_r in sec of $C_{1\text{-}2}$ in nanofd and τ_r in nanosec.

From Eq. (3), it is obvious that crosstalk is a phenomena which monotonically increase with frequency. At the worst, crosstalk could reach 0dB, i.e., the peak voltage in victim equals the peak culprit voltage (Figure 2). If the frequency where $R_{vict} \times C_{2w} = 1$ occurs before the 0dB asymptote, the crosstalk remains *clamped* to a value equal to the capacitive divider $C_{1\text{-}2}/(C_{1\text{-}2} + C_2)$.

For the mutual inductance contributor, the voltage induced *longitudinally* in the victim is

$$V_V = M_{1\text{-}2}\,\omega I_{culprit} = M_{1\text{-}2}\,\omega\frac{V_C}{Z_{C1}} \quad (5)$$

So the inductive contributor expresses as:

$$\text{Xtalk}_{(ind)} = \frac{M_{1\text{-}2}\,\omega}{Z_{cl}} \quad (6)$$

Here again it is a coupling increasing with frequency. In theory, there is no reason why the inductive crosstalk should not become greater than 0dB, i.e., the victim induced voltage being larger than the culprit line voltage, like in a step-up transformer. However, in PCB configurations, the two circuits are sufficiently similar and, being *one-turn* coupled loops, the coupling factor never exceeds unity.

Although Inductive Coupling Exists, Capacitive Coupling Generally Predominates

Since we know that 2 contributors participate to crosstalk, it seems necessary to compute them separately, then retain the larger of the two, or their combination. Hopefully in PCB's, a rough estimate can be made without this complexity.

A look at equations (3) and (6) shows that for a given geometry, *capacitive crosstalk increases with large victim impedances,* and *inductive crosstalk increases with low culprit impedances.* The take-over of capacitive coupling over the inductive one occurs for circuit loads ≅ 100Ω (very exactly when loading equals Z_0 of the line). Also the wire-to-wire capacitance is aggravated by ε_r of the epoxy glass while mutual inductance stays equal to its free-air value.

For these reasons, unless the culprit circuit is a very low impedance (< 30 Ohms), a *quick crosstalk estimate can use the capacitance coupling only.*

The table of Figure 3 gives this crosstalk per centimeter length of parallel runs on epoxy glass, for victim impedances of 100Ω and several printed traces dimensions (1 oz base material assumed, i.e., trace ≅ 20 micrometers thick). The left vertical entry is scaled in frequencies of the culprit signal while the right entry is scaled in corresponding rise times using the relation $F = 1/\pi\tau_r$ (if fall time is faster than rise time, the shortest of the two should be used).

The procedure to apply the table is the following:

- Select the geometry corresponding to the culprit-victim cross section
- Define culprit critical frequency or rise time
- Find the corresponding crosstalk per unit length (cm)
- Apply length correction

$$20 \operatorname{Log}_{10} \ell_{cm}$$

- Apply impedance correction if $Z_{victim} \neq 100\Omega$ by computing

$$20 \operatorname{Log}_{10} \frac{Z_{victim}}{100\Omega}$$

Example

Two traces have a 10cm parallel trip.
Trace width W = 20 mils
Trace separation S = 20 mils
No ground plane
Culprit = Schottky Logic, 4 volts swing, 3 nanosec rise time

Victim = Schottky Logic, Noise margin typical 1 volt, wst case 0.4 volts (input *low*)

Victim impedances = one gate *low* output impedance (about 50Ω) on driving side, parallel with one gate *low* input impedance (about 200Ω) on load side so total victim impedance = 40Ω

$$\text{Xtalk} = -38\text{ dB(table value)} + 20\text{ Log }10_{cm} \text{ (length correction)} + 20\text{ Log}\frac{40}{100} \text{ (imp. correction)}$$

$$\text{Xtalk} = -38\text{ dB} + 20\text{ dB} + (-8\text{ dB}) = -26\text{ dB} \cong 5\%.$$

$$\text{Vvictim} = 4\text{ volts} \times .05 = 200\text{mV}$$

This is about one half of the worst case signal line noise margin, so normally no problem can be expected. However, if several adjacent culprit lines, or a fan-out of several devices are running close to the victim trace, capacitive crosstalks may add up to upset the threshold of the receiving gate.

Apply Stop Band Rejection For Crosstalk Between Different IC Technologies

When different technologies are involved between the crosstalk source and victim, for instance ECL and TTL, or TTL and CMOS, or even analog devices, bandwidth limitation has to be checked. If the frequency F_{EMI} of the culprit is larger than the cut-off frequency F_{co} of the victim, a relaxation equal to 20 $\text{Log}_{10}\ F_{CO} / F_{EMI}$ should be applied to the computed crosstalk. In time domain terms, one should compute the same relaxation per 20 Log_{10} Victim rise time/Culprit rise time. This accounts for the fact that noise immunity of the victim device improves when the pulse width of the noise stimuli becomes shorter than the minimum turn-out time of this IC family. The slope of this stop-band is generally given by the gate input capacitance, i.e., 6 dB octave or 20 dB per decade.

How Geometry and Impedances Influence Crosstalk

There is an enormous amount of parameters which influence crosstalk, like trace thickness, trace self inductance, fringing capacitance in the air, mismatch factor in culprit and victim circuits.

Figure 4, summarizes the circuit parameter's dependency for crosstalk.

Simple Layout Rules Can Reduce Crosstalk at No, or Moderate, Added Cost

Several simple and easy-to-remember design guidelines can end up with a trouble-free PC board layout, rather than the *wait and see if it works* strategy:

(A) Control the ratio length/separation of parallel traces. One centimeter of parallel runs spaced by 10 mils (0.25mm) can be more fatal than 10 cm runs with 1 cm separation.

(B) Preferably run the signals *above* their ground plane or return trace, rather than co-planar layout.

(C) Strictly ban parallel runs of high speed Logic (TTL, LS, ECL) with low level analog circuits.

(D) When (B) and (C) cannot be achieved, consider a grounded *guard* trace between culprit and victim wiring.

(E) When long parallel runs are unavoidable, (example: motherboards), consider a grounded *guard ring:* a double trace surrounding the culprit (or victim) trace. This has proven very useful with long clock runs.

(F) Multilayers do include, by design,

Parameter	Crosstalk
Trace Spacing Increase	↘
Trace-To-GND (or return trace) Spacing Increase	↗
Larger ε_r	↗
Z victim LARGER	↗
Z culprit SMALLER	No influence on capacitive coupling. But may cause inductive coupling to show up.
Length of Parallel Run Increase*	↗

Note* When applying length correction add-on, besure to clamp to $\ell \leq \lambda/4$ of EMI frequency.

Figure 4. *Geometry and Impedance* vs. Crosstalk

the advantage of rule (B). But when 2 signal layers are sandwiched together without an intermediate ground layer, for cost saving, the crosstalk can be even worse than with a single layer, double sided PCB because victim and culprit traces are coupled like parallel plates, embedded in a high ε_r media.

And once you have checked your layout, do not overlook these two outsiders: the IC socket and the PCB connector. About the socket, there is not much you can do: it introduces an in situ parasitic coupling of 3pF or more, pin-to-pin. Hopefully, IC devices are generally characterized using some socket anyway and they are deemed to tolerate this coupling. The PCB connector can ruin the efforts of a sound PCB layout if care is not taken for pin assignment. Avoid assigning fast signal pins next to susceptible ones, and at least alternate signal-ground-signal-ground—this will preserve the characteristic impedance, too, if needed.

Conclusion

This short description does not pretend to encompass the whole subjec' Tons of papers have been written on this specific matter. The Bibliography on the calculation of the trace edge capacitance only would occupy a full shelf in a respectable library, notwithstanding that they give often different results, depending on the degree of sophistication of their mathematical or empirical derived nature.

But the user can rely on the Table of Figure 3 for rough prediction within 6 dB confidence, which is more than adequate in usual EMI calculations.

A good practice is to *allocate, at the beginning of a design,* a certain budget for crosstalk. For instance, no more than –20 dB (10% max) of the logic immunity threshold, for each culprit-victim team.

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Author's Background

Michel Mardiguian, is a Graduate Electro-Mechanical Engineer. He started in Electronic Packaging of Aircrafts, then in Computers Packaging at IBM R & D Laboratory. He entered the EMC area in 1974 and enjoys design and field experience of problems in the computer/telecommunications area. For four years he has been an active member of the CISPR (EMC studying body of the IEC) group on computer interference. He is Director of Training at Don White Consultants, Inc., a well-known firm which conducts seminars and consulting on Electro Magnetic Compatibility.

Insertion Loss Testing Of Common Core Filters

Filters are an important element in controlling EMC/EMI. This article describes the specification and testing of these devices.

by Kenneth Sanders
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Electronics Division
Rancho Dominguez, CA

Introduction

The last several years have seen a dramatic change in the composition of the EMI (Electromagnetic Interference) filter marketplace. Increased understanding of EMI (seperating interference into common mode and differential mode signals) coupled with changes from regulatory safety agencies and revisions to Mil-Std 461 limiting ground currents, have led to the development of a new configuration in filters, the common core or common mode filter. Also testing specifications have not kept pace with advances in technology, specifically and most notably Mil-Std 220, the specification for insertion loss testing.

Common Mode and Differential Mode Noise

To gain an understanding of what is required for common core filter insertion loss testing, a simple explanation of common mode and differential mode noise, as well as, the functioning of a common core filter is presented. The discussion that follows is stated for a two phase three wire system, but is easily extended to three phase, three and four wire systems.

Common mode noise is any noise signal that is in phase on each of the two above ground lines with respect to ground. Differential mode noise is a noise signal that is out of phase on the two above ground lines. If the 60Hz line voltage were noise, it would be differential mode. Screen room tests have demonstrated that the majority of noise below 1MHz is common mode and above 1MHz is differential mode noise.

Filter Testing

A simple common core filter is illustrated in Figure 1. The two C_1 capacitors are small in value to limit ground current and, as frequencies increase, become effective in shunting both common mode and differential mode noise to ground. The two C_2 capacitors are generally one to two magnitudes larger in value than the C_1 capacitors. They are effective only for differential mode noise as common mode noise signals would be, by definition, in phase across C_2 and thereby present no potential difference. The L_1 inductor, also called common core, is two windings on a single core. The windings are connected in a manner that the power line currents which tend to saturate a core, cancel each other and the net effective saturating current is zero. The L_1 inductor offers a high impedance to common mode signals up to a frequency where the coupling between windings falls off and leakage inductance becomes a factor. Above this frequency, L_1 becomes effective against

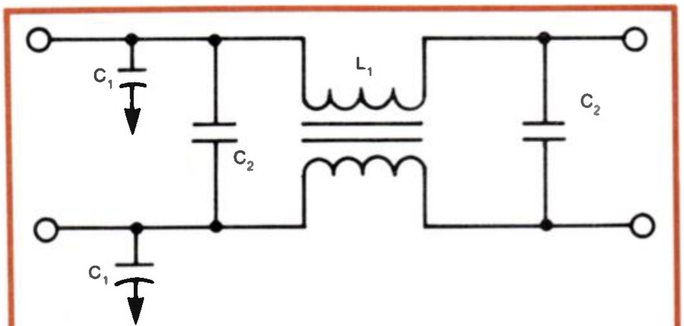


Figure 1. Typical Common Core Filter Schematic.

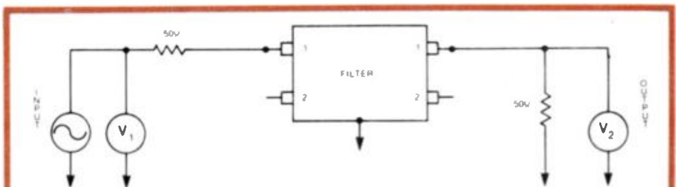


Figure 2. MIL-STD 220 Insertion Loss Test (Unused Terminals Ungrounded).

differential mode signals and less effective against common mode signals.

Mil-Std 220 insertion loss testing is not adequate to test a common core filter. The test configurations of Figure 2 and Figure 3 only inject a differential mode signal and therefore do not test for common mode signals at all. Mil-Std 220 allows for testing with unused terminals grounded or ungrounded and using the worst case. In Figure 2 only one C_1 capacitor and L_1 are effective in filtering. In Figure 3 one C_1 and both C_2 capacitors are doing filtering, but L_1 is shorted out thru terminal 2 being shorted to ground. At frequencies above L_1 winding coupling failure, L_1 becomes effective, but at only a fraction of it's original inductance.

The test circuit depicted in Figure 4 is being specified by one user of common core filters. It has most of the same disadvantages of Mil-Std 220. This method only partially shorts L_1 with 100 ohms limiting the maximum impedance of L_1 to 100 ohms. This method almost totally negates any value of the C_2 capacitors.

Insertion loss testing as depicted in Figure 5 can be accurately utilized for differential mode testing as shown. Common mode noise testing can be achieved by clamping the injection and current probes around two or three lines simultaneously. Load and source impedances are reasonably simulated, eliminating the classic argument against Mil-Std 220 testing utilizing 50 ohms. While this method is undoubtedly the most accurate, there are significant cost safety considerations for the filter manufacturer. There is the potential for operator hazard working with line voltages and currents. Test set-up time is extensive. Test time per frequency is extensive. Test readings in the 5 to 50MHz range may be difficult to obtain and above 50MHz, impossible due to line impedances and frequency/power limitations of the injection probe.

Figures 6 and 7 represent an extension of Mil-Std 220 test methods to completely test common core filters for both common mode and differential mode signals. The classic arguments against Mil-Std 220 testing in that 50 ohms does not accurately represent real world performance of a filter are still applicable. This can be easily overcome by writing filter specifications that contain both schematics and nominal element values of the filter. Elements values that are derived from testing to meet a specified conducted radiation or susceptability requirement.

Conclusion

Wide bandwidth balun transformers shown in Figure 7 are readily commercially obtainable. The test method proposed in Figure 6 and 7 can be readily adapted for full load testing by insertion of buffer networks, etc. Standardization in insertion loss test methods for common core filters is necessary to realize both cost effective and repeatable test results. □

Author's Background

Kenneth Sanders received his formal education in engineering at the University of California at Los Angeles. He has been with Genisco Technology Corporation for eighteen years as a design engineer, quality manager, and production manager. He currently is product manager for the filter product line of Genisco.

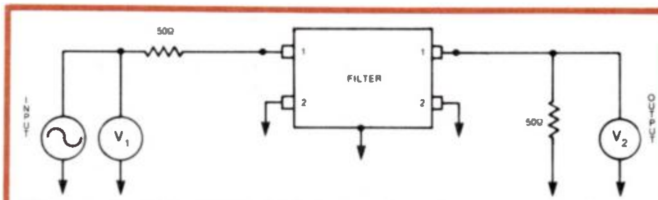


Figure 3. MIL-STD 220 Insertion Loss Test (Unused Terminals Grounded).

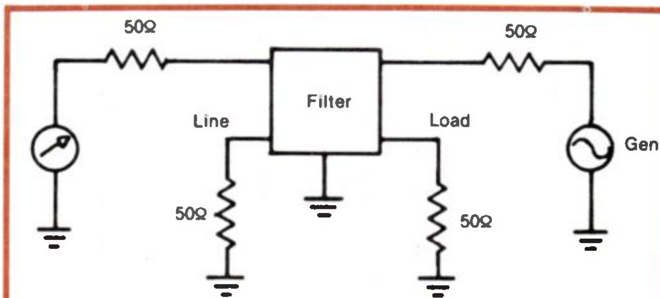


Figure 4. Proposed Test Circuit.

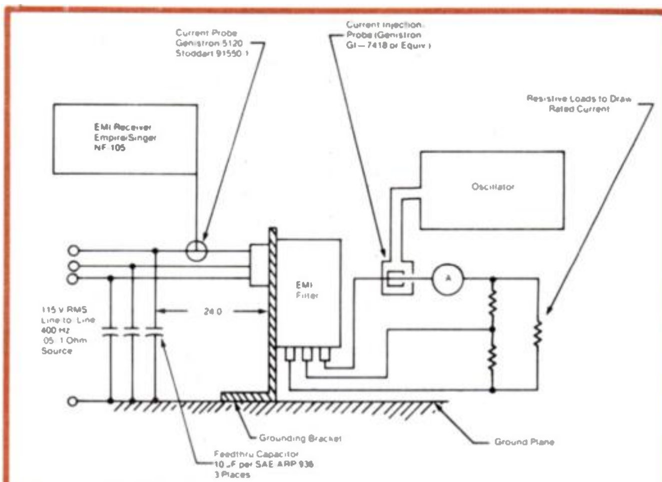


Figure 5. Ideal Insertion Loss Test.

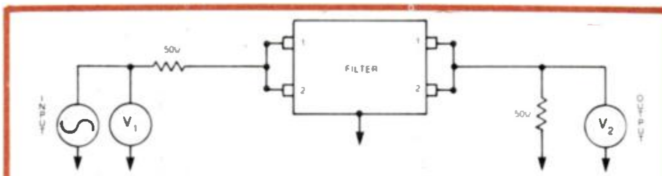


Figure 6. Simplified Common Mode Insertion Loss Test.

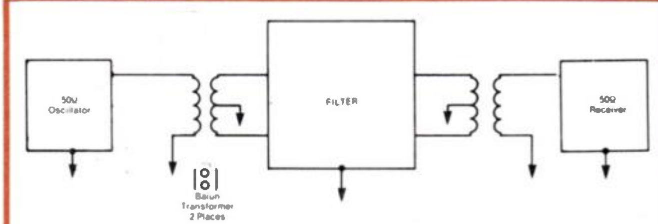


Figure 7. Simplified Differential Mode Insertion Loss Test.

A Binary Stepped Transmission Line

Binary related line lengths simplify testing procedure for high powered linear amplifiers.

By Roderick K. Blocksome
Rockwell International
Collins Telecommunications Product Division
Cedar Rapids, Iowa

The Problem

The design and development of high frequency linear power amplifiers involve specifications which describe the amplifier performance at various VSWR's referenced to a nominal output load. For example, the output tank circuit may be specified to match a 3:1 (or less) VSWR relative to 50 ohms. A broadband solid state PA may specify rated output power into loads presenting a VSWR of 1.3:1 or less. Additionally, the amplifier stability must be insured at higher VSWR's. How does one test to such specifications over the wide range of impedances of a given VSWR circle and over the frequency range of the PA? A common method used at HF (1.6 to 30 MHz) requires an adjustable matching network and a 50Ω dummy load. A reversible L-network, with sufficient tuning range, connected to a standard 50 ohm resistive load will transform the load to the desired complex impedance on the VSWR circle at the frequency of interest.

Amplifier testing at a moderate number of impedance points at selected frequencies and VSWR's can become a very time consuming task. Each point must be calculated, then with the aid of an impedance measuring instrument, the L-network adjusted, then connected to the amplifier under test and the measurement made. The time required by this method can be reduced somewhat by making up several lengths of transmission line which can be inserted between a desired VSWR load and the unit under test. Since the line electrical lengths are different at different frequencies, data taking at a constant complex impedance for different frequencies becomes a problem.

Microwave engineers use a very handy device commonly called a "sliding load" or "trombone line section". It is a low loss transmission line whose length may be quite easily changed by sliding the line in telescoping fashion. Such lines, whose length may be changed over a half wavelength, are quite manageable in the laboratory. But for testing at 1.6 MHz we would have a "trombone line section" several hundred feet long! A HF "trombone line section" could be approximated by a large number of small discrete lengths of line and a switching arrangement to connect as many lengths in series as desired. The line length would not be continously variable, but if each elemental line length were short compared to a half wavelength line at the highest test frequency, then several impedance points could be obtained at a given VSWR. For example, if the elemental line lengths were 20 electrical degrees long at 30 MHz, then 9 different impedance points on a Smith Chart VSWR circle could be transformed from a given VSWR load. At lower frequencies, more points correspondingly closer could be realized since the elemental line lengths become electrically shorter. However, the total number of elemental line sections required to achieve ½ λ (180° electrical degrees) becomes quite large with a corresponding increase in switching complexity. How can we then build a switched variable transmission line with reasonable size, cost, and complexity?

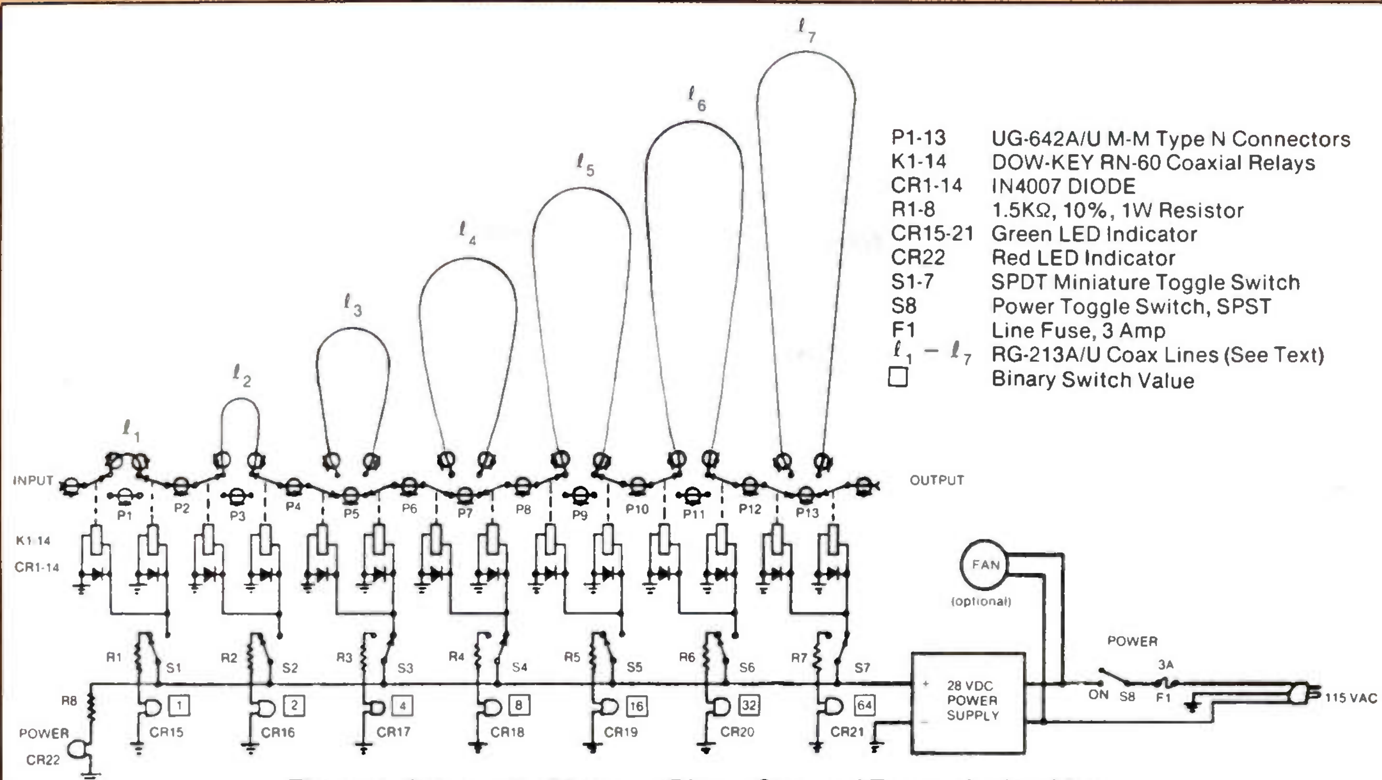


Figure 1. Schematic Diagram, Binary Stepped Transmission Line.

The Solution

The answer lies in making the "elemental" line lengths binarily related. Thus any desired total line length can be achieved in length increments equal to the smallest line length that can physically be implemented. I have designed such a device for use between 1.6 and 30 MHz. There are many ways of implementing a binary stepped transmission line within desired frequency and power requirements. A brief description of a 1 KW HF unit which was designed and built may prove helpful to others wishing to build similar units.

Circuit Description

A schematic of a Binary Stepped Transmission Line is shown in Figure 1. The unit was designed to operate up to 1000 watts, on a 50 ohm system and at frequencies between 1.6 and 30 MHz. The coaxial relay selected was the Dow-Key RN-60 SPDT with type N connectors. It was selected because of good reliability, low cost, and adequate impedance characteristics. RG 213 A/U coaxial cable was selected for the transmission lines. The velocity factor of RG-213A/U is given as 0.66 which means approximately 202 feet of cable is required to achieve ½ λ at 1.6 MHz. The smallest length of line used was 1.59 feet. This was about the shortest length that could be physically connected between the relay connectors, and was arrived at by calculating back from the maximum desired line length. The calculated lengths are

ℓ_1 = 1.59 ft.
ℓ_2 = 3.17 ft.
ℓ_3 = 6.34 ft.
ℓ_4 = 12.68 ft.
ℓ_5 = 25.37 ft.
ℓ_6 = 50.74 ft.
ℓ_7 = 101.48 ft.

Packaging 14 coaxial relays and 200 feet of RG-213A/U cable into a convenient size for general purpose lab use is not difficult if the relays are connected and stacked in a vertical zig-zag manner. The coaxial cables are dressed around the perimeter of a chassis box measuring 17-in wide × 9.75-in high × 20-in deep. Type N double male coaxial fittings are used to interconnect all relays to minimize line length through the relays. Each pair of relays at a switch point are controlled by toggle switches on the front panel. Each switch is labeled with its binary bit weight. This requires the user to do some mental addition to find the actual step number. One could readily design a relay control circuit using digital IC's and seven segment displays if the toggle switch system is too inconvenient. The first relay, K1, is mounted with its common terminal exposed at the front panel providing the input connection. The last relay, K14, is mounted in a vertical manner such that its common terminal is exposed through the top cover of the chassis to provide the output connection. This somewhat unusual arrangement precludes having internal relay interconnecting cables which would add to the inherent line length of the relays and allows the connection of loads other than 50 ohms directly and allows the connector. A fan mounted in the chassis provides cooling air to these loads. A 28 volt DC power supply was incorporated in the unit to power the relays. The cover picture shows the interior of the completed unit.

VSWR Loads

The load impedance required to produce various VSWR's may easily be obtained either of two ways. The easiest is to connect two, three, or four commercial 50-ohm dummy loads in parallel using "Tee" connectors to produce 25, 16.67, and 12.5 ohm loads for tests at 2:1, 3:1, and 4:1 VSWR. If other intermediate VSWR's are desired, large wattage carborundum type resistors are available to construct a load suitable for operation up to 30 MHz. Experiment with resistors and an impedance measuring instrument to obtain the desired resistive load with negligible reactance at

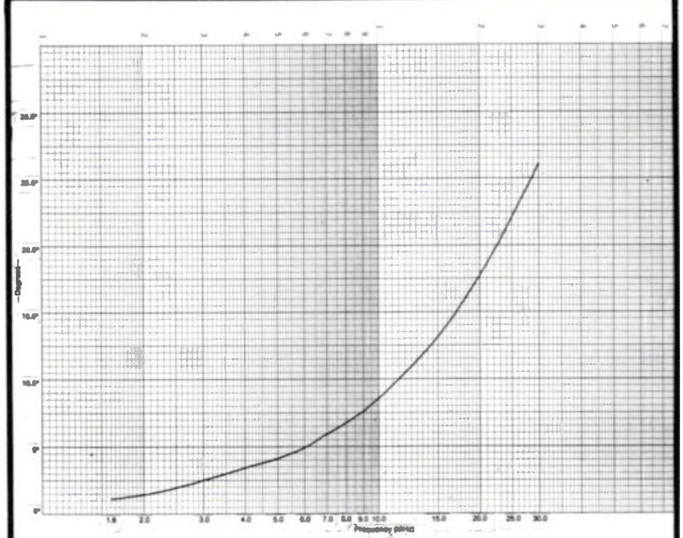


Figure 2. Line Length in Electrical Degrees per Unit Step Versus Frequency in MHz.

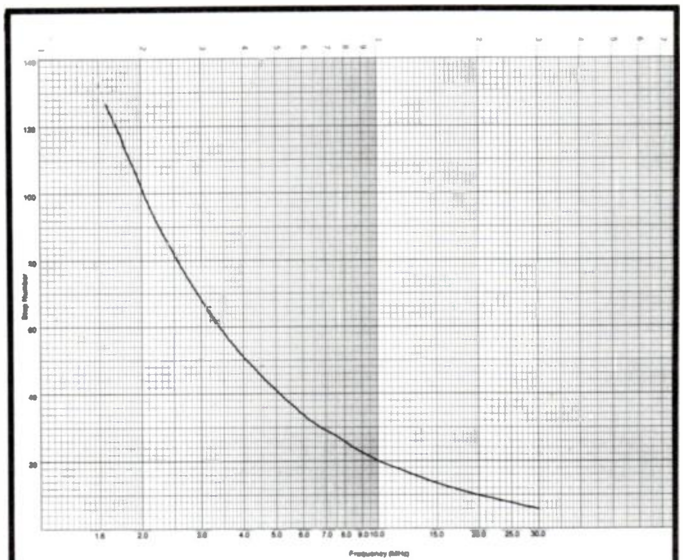


Figure 3. Step Number for 180° Rotation from Step 0 Versus Frequency in MHz.

30 MHz. Another possible source would be custom load impedances from manufacturers of dummy loads. Perhaps one could purchase otherwise rejected resistor elements in the range of 40 to 60 ohms for low VSWR testing.

An alternative method of obtaining low VSWR resistive loads is to use attenuators terminated into either an open circuit or short circuit. Such attenuators are commonly available in lower power ratings. Table 1 lists VSWR which may be obtained from common attenuator values. The formula for calculating the values in Table 1 is provided for those desiring other values.

Using the Binary Stepped Line

When using the binary stepped transmission line, it is convenient to know the number of electrical degrees of line produced, per unit step, at the frequency of interest. The graph shown in Figure 2, kept with the equipment, will aid the user. Line losses will introduce an error in actual VSWR (VSWR reduction with increasing line length and line loss). For this reason it is a good practice to always use line lengths of ½ λ or less even though longer lines are possible at higher frequencies. Remember, impedance transformation on lines up to a ½ λ cover a full VSWR circle on the Smith Chart. A second graph shown in Figure 3 gives the step number versus frequency to produce ½ λ of line. As mentioned previously, VSWR will decrease slightly as one inserts longer and longer line lengths between the load and the device under test. This loss is also a function of the VSWR on the line. This should not be a problem for most applications if the maximum VSWR is held to 5:1. In tests which require greater impedance or VSWR accuracy, it is a simple matter to set up the test with the binary stepped transmission line and an HP-4815A RF Vector Impedance meter to read the impedance. The VSWR for that impedance can readily be found by solving the following equation on a handheld calculator. An HP-25 program is given in Table 2.

Table 1. VSWR of Various Attenuators

Attenuator Value	VSWR
1 dB	8:72:1
2 dB	4.42:1
3 dB	3.01:1
4 dB	2.33:1
5 dB	1.93:1
6 dB	1.67:1
7 dB	1.50:1
8 dB	1.38:1
9 dB	1.29:1
10 dB	1.22:1
20 dB	1.02:1

Formula for calculating VSWR of an attenuator terminated in either an open or short circuit.

$$\text{VSWR} = \frac{10^{.1A} + 1}{10^{.1A} - 1}$$

where A = Attenuation in dB

$$\text{VSWR} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \qquad \Gamma = \frac{Z_L - Z_O}{Z_L + Z_O}$$

where: $Z_L = R_L + jX_L$ load impedance

Z_O = characteristic line impedance

assuming Z_O is real, solving for $|\Gamma|$:

$$|\Gamma| = \left[\left\{ \frac{\left(\frac{R_L}{Z_O} - 1\right)\left(\frac{R_L}{Z_O} + 1\right) + \left(\frac{X_L}{Z_O}\right)^2}{\left(\frac{R_L}{Z_O} + 1\right)^2 + \left(\frac{X_L}{Z_O}\right)^2} \right\}^2 + \left\{ \frac{2\frac{X_L}{Z_O}}{\left(\frac{R_L}{Z_O} + 1\right)^2 + \left(\frac{X_L}{Z_O}\right)^2} \right\}^2 \right]^{1/2}$$

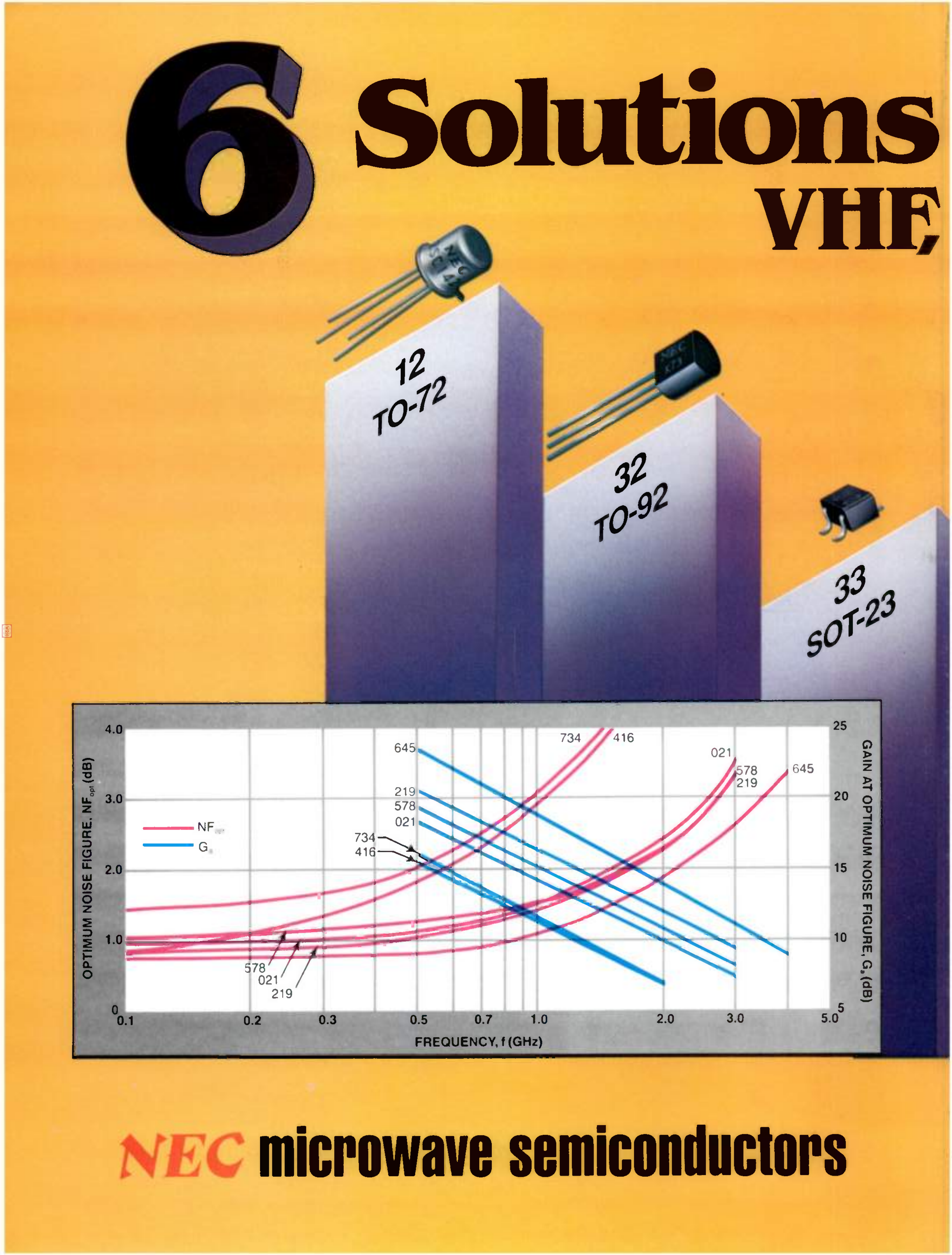
Increasing electrical line length per step with increasing frequency permits fewer impedance points to be accessable around the VSWR circle. A "trombone line" borrowed from the UHF lab and inserted in series with the binary stepped line will allow access to these intermediate impedance points.

Applications

Applications of the binary stepped transmission line in HF department work are numerous. Some uses that it has been put to in our HF lab are:

1) Tuning range evaluation of transmitter output networks and antenna couplers.

(Continued on page 29.)



6 Solutions
VHF,
12
TO-72
32
TO-92
33
SOT-23
NEC
OPTIMUM NOISE FIGURE, NF_opt (dB)
GAIN AT OPTIMUM NOISE FIGURE, G_a (dB)
FREQUENCY, f (GHz)
NF_opt
G_a
4.0
3.0
2.0
1.0
0
0.1
0.2
0.3
0.5
0.7
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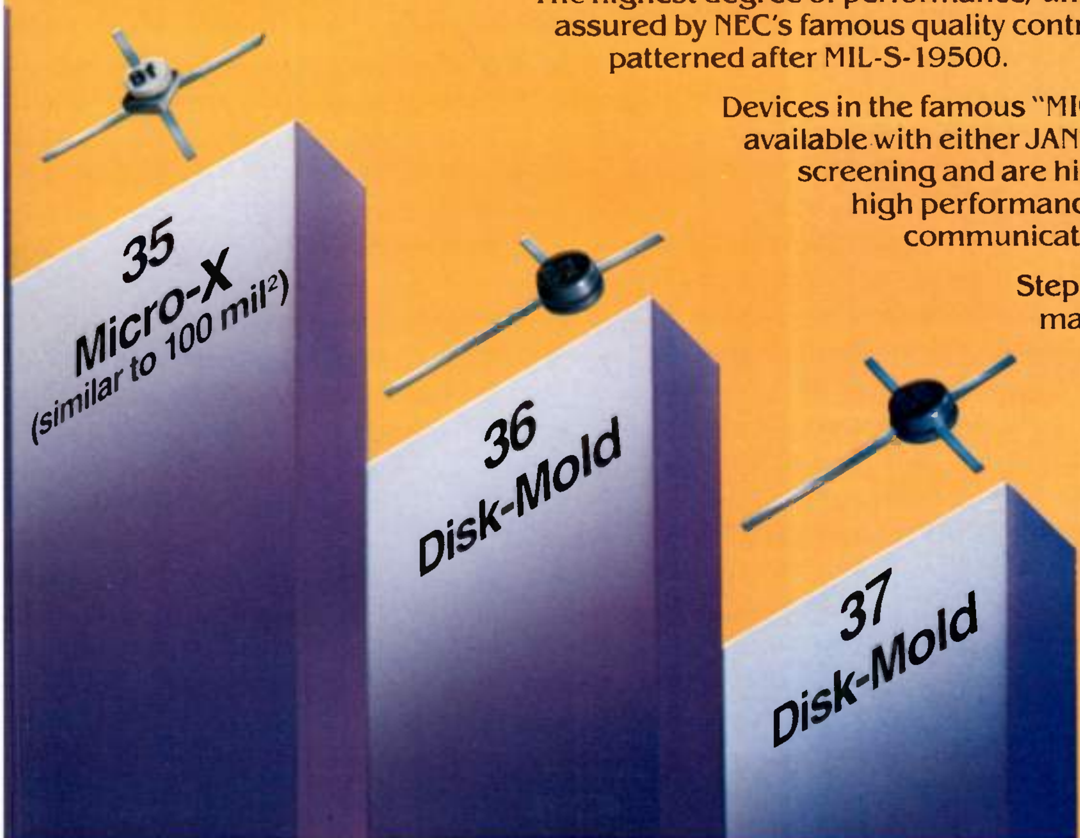

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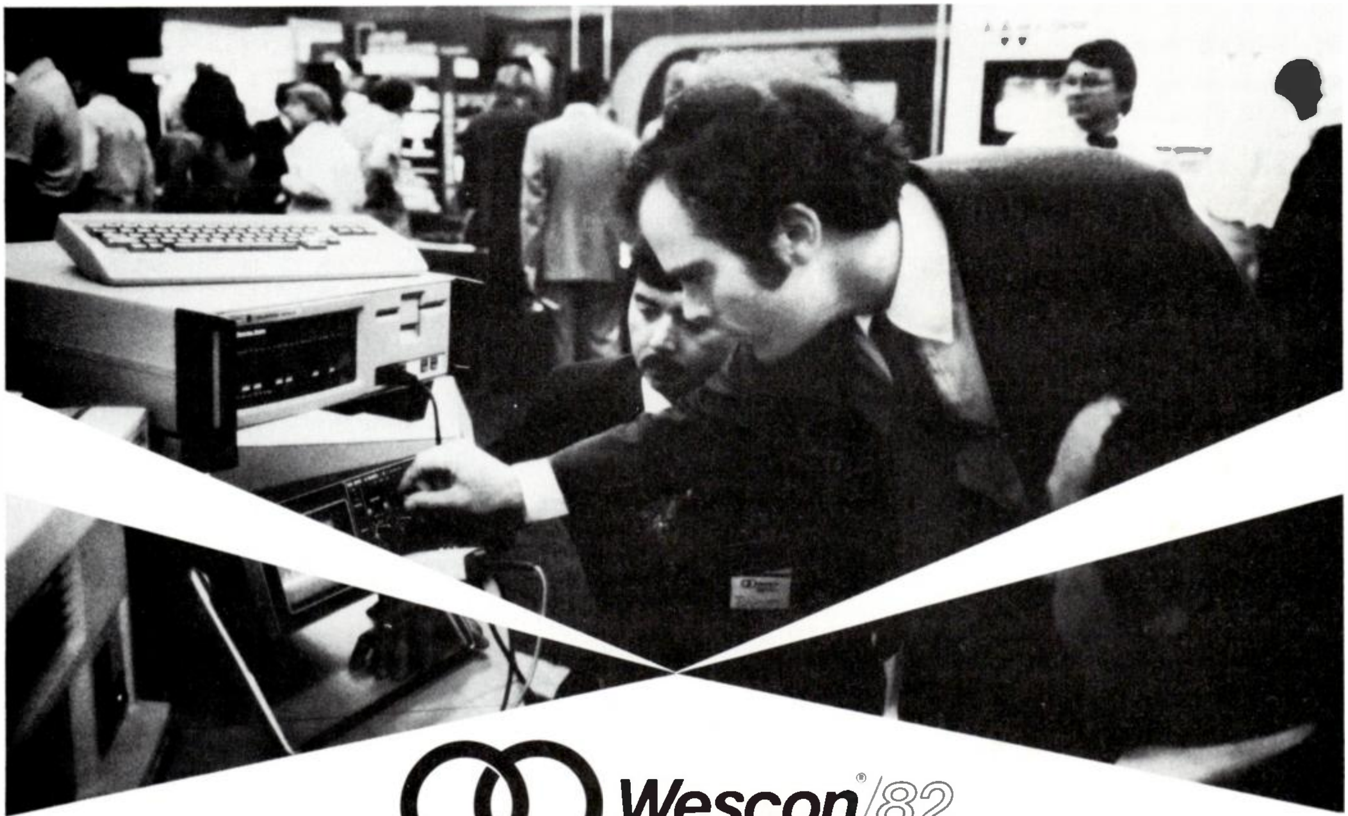

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Table 2.
HP-25 Program to Solve for VSWR, Given Impedance.

STEP	PROGRAM	STEP	PROGRAM
1	RCL 0	23	+
2	÷	24	X ↔ Y
3	STO 1	25	÷
4	STO 2	26	g X^2
5	1	27	STO 5
6	STO − 1	28	RCL 3
7	STO + 2	29	2
8	R↓	30	X
9	R↓	31	RCL 4
10	RCL 0	32	÷
11	÷	33	g X^2
12	STO 3	24	RCL 5
13	g X^2	35	+
14	RCL 2	36	f $\sqrt{X}$
15	g X^2	37	STO 6
16	+	38	1
17	STO 4	39	+
18	RCL 1	40	RCL 6
19	RCL 2	41	CHS
20	X	42	1
21	RCL 3	43	+
22	g X^2	44	÷

STORE Z_0 in Register 0

a) Input Data for Z_L in Rectangular form:
  key in X_L
  ENTER
  key in R_L
  ENTER
  R/S

b) Input Data for Z_L in Polar form:
  key in angle 0 (degrees)
  ENTER
  key in magnitude |Z| (ohms)
  f → R
  R/S

(Continued from page 24.)

2) Evaluation of solid state PA dissipation protection controls under VSWR conditions.
3) Evaluation of filter input harmonic impedances and their relationship to solid state power amplifier performance.
4) Solid State PA stability tests under VSWR conditions.

Within the limitations of the impedance transformation accuracy described above, the binary stepped transmission line provides a means of rapidly transforming any known or given impedance to any other desired impedance on essentially the same VSWR circle. Changes in test frequencies are likewise handled rapidly and efficiently.

Acknowledgment

The author wishes to acknowledge the contribution of Forrest Arnold in packaging design and construction of a working model. □

Bandpass Filters With Self-Equalized Group Delay: Part 2

Part I dealt with the background, practical and theoretical considerations dealing with this design method. Part II gives the details of the filter design.

By Robert W. Sellers
Harris Corp., GESD
Melbourne, Florida

Filter Design Method

The procedure to be described is admittedly not mathematically exact, but it is sufficient to calculate correct capacitor values to within a fraction of 1%. The calculated values may be easily adjusted via either the computer program or in the case of actual circuit construction, obtained by measurement on a well calibrated capacitance bridge, and trimmed slightly in final alignment of the network.

As mentioned previously, the procedure to be described is based on a design with four inband resonators and two correction poles. For bandwidths less than about 25% of center frequency, a pair of series tuned compensator circuits are used to improve the group delay even further in the transition regions. These compensator circuits are due to Lerner and comprise the means by which the reactive residual components of the network are effectively included as a part of the filter loads. For the wider bandwidth (36 MHz) filters, the low Q necessary in these compensators produce excessive non-symmetry in the amplitude response. For the 36 MHz filter, the group delay ears are produced by the pair of out-of-band zeros which are used to adjust the out of band attenuation.

It should be realized that there are two basic choices for implementation of the filter to be described shortly. For example, the filter may be designed using linear frequency spacing across the passband; in this case, exponential component sizing is necessary for the networks. The design is equally valid for the case where each inband resonator uses the same magnitude of capacitance and the frequency spacing is made exponential across the passband. The case of linear frequency spacing is selected for the design example.

Example

The example will now be presented via a 14 step design procedure of a filter with the following characteristics, schematically represented by Figure 6.

Center frequency (f_O) = 70 MHz
Bandwidth (3 dB) = 10 MHz
Inband resonators = 4

(1) Determine the frequency spacing necessary between parallel resonances:

$$\frac{10\text{ MHz BW}}{4\text{ resonances}} = 2.5\text{ MHz} = 2\Delta F$$

(2) Calculate the actual center frequency of the resonances:

Since there is an even number of networks in the total filter (both arms), there is no single network resonance at exactly 70 MHz. The two closest (f_3 and f_4) will therefore be located at $\pm \frac{2\Delta f}{2} = \Delta f \pm 1.25$ MHz away from 70 MHz. Following Lerner, we will choose to use the linear frequency spacing. However, because we choose to do this, we must abandon the use of the same capacitor value for each of the resonators, we therefore arrive at the frequencies of the resonances as follows:

$f_2 (L_2 - C_2) = 66.25$ MHz
$f_3 (L_3 - C_3) = 68.75$ MHz
$f_4 (L_4 - C_4) = 71.25$ MHz
$f_5 (L_5 - C_5) = 73.75$ MHz

f_1 and f_6 are the corrector poles and will be nominally adjusted to be at the upper and lower band edges of the filter. The calculated frequency for these resonances would be

65 and 75 MHz respectively in the Lerner design. The Lerner method requires that the inband capacitors each be the same magnitude and the corrector capacitances twice that value. We will later arrive at a nominal value for these corrector capacitances which are $\cong \pi/2$ rather than 2 times the inband capacitor values. This in turn requires that the frequencies for the corrector resonances be placed $\cong$ 20% (of $2 \Delta f$) out of band. The correctors will be determined in the final steps of this procedure.

(3) Place the resonances from the lowest inband frequency (f_2) to the highest inband frequency (f_5) in alternate arms of the lattice. At this point, f_2 and f_4 are in say, arm B, while f_3 and f_5 reside in arm A. A space, f_1, is reserved for the low frequency corrector pole in arm A, and likewise, a space for corrector f_6 is reserved in arm B.

(4) Determine the desired impedance level for the filter. Remember that as in Figure 3, the filter sees a total external resistance of R, which is the sum of the source and load resistance. This step allows the freedom to choose the approximate magnitude of the capacitors to be used in the filter implementation.

$$R(\text{Filter}) = (4/\pi)(Xc) \mid f = 2\Delta f$$

where Xc is the reactance of the selected capacitor evaluated at a frequency of $2 \Delta f$. (2.5 MHz in this example) Say the desired total load resistance is 500 ohms; then C = 100 pF.

The capacitor value thus calculated is the "nominal" value to be used in the implementation. It is important to realize that a capacitor of exactly 100 pF will not actually be used. Likewise the filter source and load resistance will not be exactly 250 ohms each; herein lies the root of some of the mathematical inexactness. It will later be found, however that the effective performance of the filter will be well within predicted tolerances in implementation.

The significance of the 100 pF value which was calculated may be explained by realizing that with an even number of resonators, no network is required at the filter center frequency. If an odd number was selected for the inband resonators, it would use such a capacitor value. Further, it is customary to assume that any practical filter will only exhibit the calculated resistive component at precisely f_O. Small mismatches at inband frequencies other than f_O are tolerated in this design as well.

(5) Select an unloaded Q for the resonators. This is an important step as all the Q's should be the same. The choice of inductor style as well as capacitor type and frequency range to be covered (B W of filter) will have an impact on this selection. As with any filter, we wish to have the greatest unloaded Q possible, to minimize the filter insertion loss.

For the example filter, we need inductors in the 50 nh range to resonate the 100 pf capacitors near 70 MHz.

The Q of the inductors should be measured and a good average (practical value) employed for the calculations which follow. Say the average of this unloaded Q is Qu = 100, evaluated at 70 MHz.

$$X_C(\text{nom}) \text{ at } 70 \text{ MHz} = 22.7\Omega.$$
$$\therefore R_p = 2.27K = (Q)(X_c);$$

(7) Calculate $\propto_{nom} = \dfrac{1}{2RC}$ for the nominal values.

$$\propto_{nom} = \frac{1}{2RC} = \frac{1}{2(2.27K)(100pf)} = 2.2026 \times 10^6$$

(8) The results of step 7 is the Neper frequency we wish to scale as a constant across the band of the filter. Since the filter is symmetrical about f_O, we need only to calculate $\propto$ at the resonance above, (f_4) and below, (f_2) f_O. Both these resonances appear in arm B. Arm A will then use the same capacitor values but in reverse order.

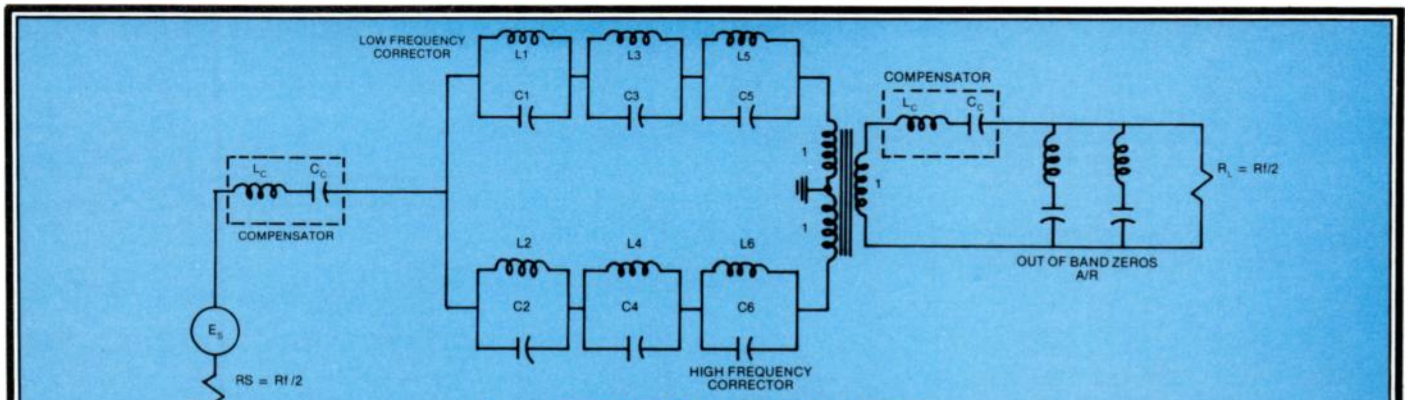


Figure 6. Filter Schematic Diagram.

$$\propto(f_2) = 2.2026 \times 10^6 \left(\frac{66.25 \text{ MHz}}{70 \text{ MHz}}\right) = 2.0846 \times 10^6$$

$$\propto(f_4) = 2.2026 \times 10^6 \left(\frac{71.25 \text{ MHz}}{70 \text{ MHz}}\right) = 2.2419 \times 10^6$$

(9) Arm B contains individual parallel resonances at f_2 and f_4 representing finite attenuation poles. It is necessary, however, that a frequency f_3, arm B contain a transmission zero. In order for arm B to be series resonant at f_3, it is necessary to employ the capacitance residual from f_2 (above resonance) and obtain the proper magnitude of inductance from f_4 (below resonance). These solutions must be simultaneously achieved while maintaining the correct $\propto$ and parallel resonances.

$$\propto(f_3) = 2.2026 \times 10^6 \left(\frac{68.75}{70}\right) = 2.163 \times 10^6$$

Note that $\propto(f_3)$ must be obtained from a series resonant circuit where $\propto = \dfrac{R'}{2L}$

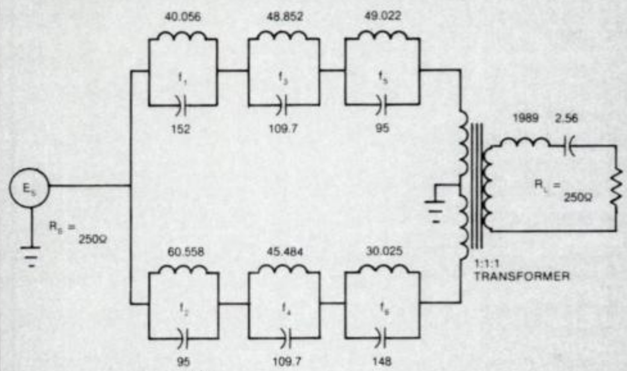
$$\text{where } R' = \frac{R_p}{1 + Q^2} = 0.227\Omega$$

we now calculate:

$$\frac{R'}{2L} = 2.163 \times 10^6; \quad \frac{1}{L} = \frac{(2)(2.163 \times 10^6)}{(0.227)}$$

$$L = 52.47 \text{ nH}$$

This is a mythical inductance. It will be used to calculate the final capacitor values necessary at f_2 and f_4 to provide the proper $\propto$. The same comment regarding mathematical



f_1 = 64.5 MHz
f_2 = 66.25 MHz
f_3 = 68.75 MHz
f_4 = 71.25 MHz
f_5 = 73.75 MHz
f_6 = 75.5 MHz

INDUCTORS IN NANOHENRIES
CAPACITORS IN PICOFARADS
Q_U OF EACH RESONANCE = 100 @ 70 MHz

Figure 7. Final Filter Schematic with Values, Bw = 10 MHz.

exactness as discussed in step 4 for the nominal capacitor applies here.

The capacitor values are calculated using the standard approximation for resonances, $\omega_o = \frac{1}{\sqrt{LC}}$ (R>>ωL)

$$66.25 \text{ MHz} = \frac{1}{(2\pi)\sqrt{52.47 \times 10^{-9} C_2}}$$

C_2 = 109.99 pf; (f_2 = 68.75 MHz)

Likewise, C_4 is found to be 95.1 pf. (f_4 = 71.25 MHz)

(10) The simple calculations above must now succumb to some logical reasoning. C_2 and C_4 were calculated based upon the ∝ we wish to be simultaneously exhibited by the *other* arm—the one containing f_3 and f_5. Since the value for ∝ should be scaled across the entire filter, and not just in one arm, the capacitors calculated to be used as C_2 and C_4 must be actually used for C_3 and C_5 respectively! This requires that the values for C_2 and C_4 be reversed to provide the exponential value symmetry for each lattice arm *and* for the filter simultaneously. With this realization, the capacitor values calculated may be placed in the filter as follows:

Arm "B" { f_2(66.25 MHz) C_2 value = 95.1 pf
f_4(71.25 MHz) C_4 value = 110 pf

Arm "A" { f_3(68.75 MHz) C_3 value = 110 pf
f_5(73.75 MHz) C_5 value = 95.1 pf

The corrector resonators, f_1 and f_6, are yet to be determined.

(11) With the capacitor values thus selected, it is a simple matter to parallel each value with an inductor to provide the proper center frequency using the resonance approximation, $\omega_o = \frac{1}{\sqrt{LC}}$

```
100  PLC  AA  SE  40.056892   152    100   70
110  CAP  BB  PA  .0000001    1000  70
120  PLC  CC  SE  48.852662   109.7    100   70
130  CAP  DD  PA  .0000001    1000   70
140  PLC  EE  SE  49.022179   95    100   70
150  CAP  FF  PA  .0000001   1000   70
160  TRF  GG  TR  1
170  CAX  AA  GG
180  PLC  HH  SE  60.558568   95.3    100   70
190  CAP  II  PA  .0000001   1000   70
200  PLC  JJ  SE  45.48455   109.7   100   70
210  CAP  KK  PA  .0000001   1000   70
220  PLC  LL  SE  30.025113   148    100   70
230  CAP  MM   PA   .0000001   1000   70
240  CAX  HH  MM
250  TRF  XX  TR  1
260  CAS  HH  XX
270  PAR  HH   AA
280  EQU  QQ  HH
290  SLC  NN  SE  1989   2.56   100   70
300  RES  OO  SE  .01
310  CAP  PP  PA  .0000001   1000   70
320  RES  RR  SE  .01
330  SLC  SS  SE  .001   1000000   1000000   70
340  CAX  NN  SS
350  PLO  NN  S7  250
360  PRI  NN  S7  250
370  END
380  60   70   1
390  71   80   1
400  END
READY.
```

Figure 8a. Compact Program for 10 MHz Bw Filter.

	S11		S21		S12		S22		
Freq.	Magnitude	Angle	dB	Angle	Magnitude	Angle	Magnitude	Angle	Delay
60.00	.92	−111	−28.33	−89.2	.038	−89.2	.96	131	32.909
61.00	.89	−128	−24.38	−103.1	.060	−103.1	.94	123	44.651
62.00	.83	−152	−19.76	−122.7	.103	−122.7	.91	113	64.729
63.00	.72	172	−14.36	−152.3	.191	−152.3	.85	97	101.694
64.00	.49	109	−8.76	160.0	.365	160.0	.69	74	161.872
65.00	.23	4	−5.06	92.7	.558	92.7	.38	46	200.003
66.00	.10	−76	−3.93	20.3	.636	20.3	.08	38	200.288
67.00	.11	−76	−3.88	−51.3	.640	−51.3	.11	129	198.008
68.00	.10	−115	−3.89	−122.9	.639	−122.9	.08	100	200.592
69.00	.07	−104	−3.85	164.7	.642	164.7	.06	−159	200.221
70.00	.07	−142	−3.83	93.2	.643	93.2	.08	−178	198.155
71.00	.03	−122	−3.83	21.5	.643	21.5	.05	−131	200.052
72.00	.06	−174	−3.88	−50.7	.640	−50.7	.11	−112	200.576
73.00	.04	52	−3.85	−122.7	.642	−122.7	.11	−107	199.314
74.00	.01	6	−3.91	165.5	.637	165.5	.11	−62	200.972
75.00	.25	−17	−5.15	92.2	.553	92.2	.39	−42	201.621
76.00	.58	−92	−9.19	28.1	.347	28.1	.72	−70	150.397
77.00	.75	−144	−14.36	−14.5	.191	−14.5	.87	−93	95.157
78.00	.83	−178	−19.04	−42.1	.112	−42.1	.93	−108	63.492
79.00	.88	158	−23.02	−61.2	.071	−61.2	.95	−118	45.486
80.00	.90	139	−26.42	−75.2	.048	−75.2	.96	−125	34.359

Polar S-Parameters in 250.0 Ohm System

Figure 8b. Computer Calculated Values for S-Parameters and Delay for Figure 8a.

The resulting values for the example are:

$$\text{Arm A}\begin{cases} C_3 = 110\text{ pf}, L_3 = 48.719\text{ nH}, Q_u = 100, f = 68.75\text{ MHz} \\ C_5 = 95.1\text{ pf}, L_5 = 49.022\text{ nH}, Q_u = 100, f = 73.75\text{ MHz} \end{cases}$$

$$\text{Arm B}\begin{cases} C_2 = 95.1\text{ pf}, L_2 = 60.749\text{ nH}, Q_u = 100, f = 66.25\text{ MHz} \\ C_4 = 110\text{ pf}, L_4 = 45.360\text{ nH}, Q_u = 100, f = 71.25\text{ MHz} \end{cases}$$

(12) The values of the corrector capacitors must be selected to achieve two simultaneous results.

(a) The magnitude of the resonator must have the desired effect of cancelling the closed inband pole in the opposite lattice arm, and

(b) The effect must exponentially vanish out of band with a damping coefficient such that it provides no overshoot and maintains the proper ∝ to insure a (resistance) transition to a low value. The termination of the filter for the transition out of band is extremely important, however at frequencies as far as 2 Δ f out of band, the termination of the filter is relatively unimportant allowing the use of zeros as mentioned earlier.

Continuing with the idea of exponential sizing for the filter capacitor values (since we have used a linear frequency spacing), we may estimate the corrector capacitor values as follows:

$$C_c\,(\text{nominal}) = \frac{1}{1 - e^{-1}}\ (95.1\text{ pf}) = 150.3\text{ pf}$$

The value 95.1 pf will be recognized as the smallest value in each arm. Since the "nominal" capacitor value which was originally calculated in step 4 was 100 pf and not 95 pf, a closer estimate for the two corrector capacitors may be found by allowing for the average difference between these values.

$$\text{We have } C_c = 150.3\text{ pF} \pm 2.5\text{ pF}$$

The larger value ≅ 152 pF would be associated with f_1, while the smaller value of 148 pF would be used with f_6.

(13) The inductances to be used are calculated as in step 11, and the final corrector values are found to be:

ARM A; $f_1 = 65$ MHz, $C_1 = 152$ pF, $L_1 = 39.44$ nh, $Q_u = 100$
ARM B; $f_6 = 75$ MHz, $C_6 = 148$ pF, $L_6 = 30.426$ nh, $Q_u = 100$

At this point it should be noted that since the corrector capacitors are not twice the value of the inband capacitors, the transmission zero formed by the corrector and its closest inband neighbor in each arm will not fall precisely on top of the pole to be cancelled in the opposite arm. This dilemma is solved by purposely tuning the correctors out of band by 20% of the 2 Δ f value, in the case of the example, this amounts to 500 kHz. The values of L_1 and L_6 are adjusted to do this, resulting in the new values for L_1 and L_6 of 40.057 and 30.025 nh, respectively.

(14) The final step is to calculate a compensation circuit which will be placed in series with the entire filter. The use of this circuit is completely due to Lerner, and is found to consist of a series LC circuit resonant as f_O. This circuit is useful to absorb the reactive components of both the filter and the source/load. Since the network is in series with the filter, rather than in parallel as it would be in the y-style Lerner filter, it is necessary that its bandwidth be greater than that of the filter. For the z-transformed filter using the capacitor values of 2X for the corrector poles, the Lerner theory reveals that the Q required of this network be 3/5 that of the filter.

The use of the exponential methods described in this recipe results in non-existent or very minimal group delay ears; therefore the effect of this network on the filter is much less.

It has been found experimentally that such a network is useful indeed for "tuning" the source and load reactances. Via use of a computer program we have determined the correct value range for these components to be non-critical and reasonably close to an inductive reactance that is 5/3 of the *total* loop resistance of the filter. For the example, when the losses due to the finite Q's of the networks is included, X_L is found to be 874 ohms corresponding to an inductor of about 2 μH. The capacitance required for 70 MHz resonance is then found to be 2.5 pF.

The Final Schematic for the filter is given in Figure 7. This schematic is used for the computer analysis shown in Figures 8 a-c. The performance of the filter as constructed is shown in Figure 9. The slight change in the values of C_3 and C_4 to 109.7 pF from the calculated values of 110 pF was a result of changing the corrector resonator center frequency slightly out of band as mentioned in Step 13.

Other Analysis

Additional analysis and work was done on 70 MHz filters with 5 MHz, 17 MHz and 36 MHz BW filters of this type. The effect of out-of-band zeros was noticed on the implementations of the 36 MHz filter. These components were not included in the 36 MHz computer program. Their addition enabled much better symmetry for the near skirts at the expense of some group delay ears. In particular, the filter implementation for the 36 MHz filter contains a total of only 8 individual LC circuits including the out-of-band zeros, resulting in the elimination of considerable circuitry over the present means of fitting this particular Intelsat mask.

The Computer Program

The analysis program will be recognized as written in COMPACT*, a very useful R.F. analysis tool for both active and passive networks. It is available under this name via G.E. timeshare, or from C.D.C. Corporation, where it is termed RFOPT. It is also available directly from Compact Engineering.

The actual programs contain lines which were used during development of the filters to experiment with the effects of non-perfect layout and components. Their effect

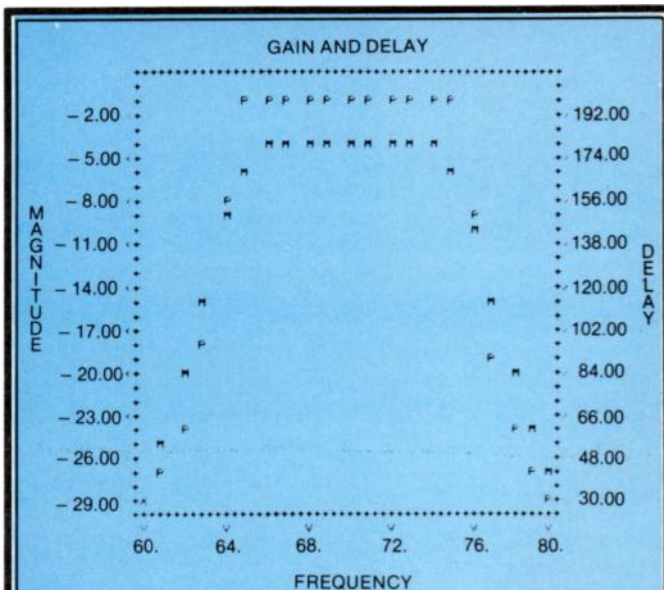


Figure 8c. Computer Generated Plot of $|S_{21}|$ (M) and Delay (P) for 10 MHz Bw Filter.

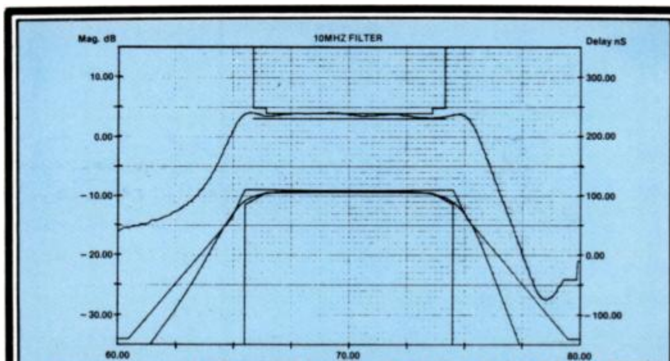


Figure 9a. Amplitude and Delay of 10 MHz BW Filter as Built. Masks are Standard Intelsat.

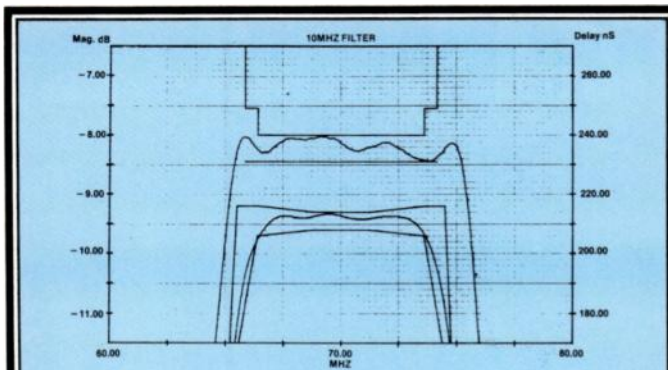


Figure 9b. Nose Characteristics of Figure 9a.

on the analysis presented is nil because the values set into them as shown for example in Figure 8a, lines 110, 130, 140, 190, 210, 230, and 300 thru 330, have no effect at the evaluation frequencies.

Breadboard Photograph

The cover of May/June issue of *R.F. Design* is a photograph of a 36 MHz BW filter. The 180° transformer is mounted beneath the p.c. card. The Motorola MWA series amplifiers, together with S.A.T. resistor networks are used to set the module gain that was needed for a particular application. *All* filter components are shown in the photo (with the exception of the transformer). □

*COMPACT information is available from: Compact Engineering, Inc., 1088 Valley View Court, Los Altos, California 94022.

Errata

In part 1 in the last issue, the following errors were noticed.

The corrected equations should be:

$$AN(\omega) = \left(1 + \frac{\omega^2 \ln 2}{N}\right)^{-N/2} \quad (6)$$

$$\phi N(\omega) = N \tan^{-1}\left(\frac{\omega\sqrt{\ln 2}}{\sqrt{N}}\right) \quad (7)$$

$$A = e - \left(\omega^2 \frac{\ln 2}{2}\right) \quad (8)$$

Also the caption under figure 2b should read "nose response" rather than "noise response".

Triple Tuned Circuits

Previous articles described double tuned coupling circuits, now a wide band triple tuned coupling circuit offering.

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

While the previously described double tuned circuits[1, 2] offer simplicity, sometimes a wide band coupling circuit with more stop band attenuation is needed. The triple tuned circuit may be the answer. It has 18 dB/octave ultimate attenuation versus 12 dB/octave for the double tuned version and is very easily implemented. The described technique has been used successfully, using standard fixed military RF chokes and ceramic capacitors, for applications requiring up to 3:1 bandwidths.

The basic technique is described by Terman[3]. It can be modified to include finite Q values of inductors. The circuit is shown in Figure 1 and equations for the modified component values are:

$$L_1 = \frac{R'}{\pi(f_2 - f_1)} \qquad (1)$$

$$C_1 = \frac{f_2 - f_1}{4\pi f_1 f_2 R'} \qquad (2)$$

$$L_2 = \frac{R_O[Q_L(f_2 - f_1) - \sqrt{f_1 f_2}]}{2\pi Q_L f_1 f_2} \qquad (3)$$

$$C_2 = \frac{1}{2\pi(f_2 - f_1)R'} \qquad (4)$$

where
f_1 and f_2 are the passband limits $w = (f_1 + f_2)/2$
and
R' is R_O and $Q_L \omega L_2$ in parallel $R' = R_O \parallel Q_L w L_2$

Terman implies that $R_O = R_i$, but it will be shown that good performance can be obtained when that is not the case.

The HP-41C program of Table I calculates the values of L_1, C_1, L_2 and C_2 using equations (1), (2), (3), and (4).

The program of Table II calculates the frequency response in dB (20 log e_O/e_i).

The following may serve as an example:

R_O = 75 ohms
R_1 = 25 ohms
f_1 = 20 KHz
f_2 = 60 KHz
Q_L = 50

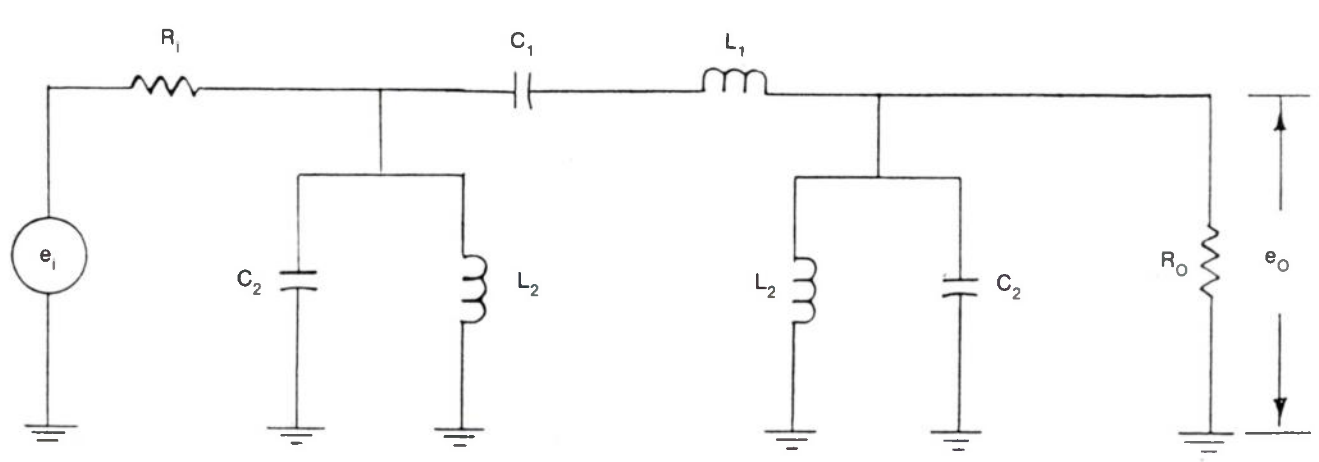


Figure 1. Triple Tuned Circuit

The calculated circuit values are:

$L_2 = 3.91 \times 10^{-4}$ H
$L_1 = 5.88 \times 10^{-4}$ H
$C_2 = 5.39 \times 10^{-8}$ F
$C_1 = 3.59 \times 10^{-8}$ F

The frequency response can then be calculated and the plot is shown in Figure 2. (Use SF 00 for the PRPLOT routine).

The 3 dB points are at 22 KHz and 56 KHz. This can be compensated for, if necessary, by changing f_1 and f_2 accordingly and recalculating the values of L_1, L_2, C_1 and C_2. The calculated response at geometric center is −2.8 dB. The actual value, using lossless inductors should be −2.5 dB (20 log 75/100); thus about 0.3 dB is lost in the inductors due to their finite Q. ☐

References

1. A.B. Przedpelski, "Double Tuned Circuits", r.f. design, Jan/Feb 82.
2. A.B. Przedpelski, "Low Impedance Double Tuned Circuit", r.f. design, May/June 82.
3. F.E. Terman, "Radio Engineers' Handbook", McGraw-Hill Book Co., NY, 1943, p. 231.

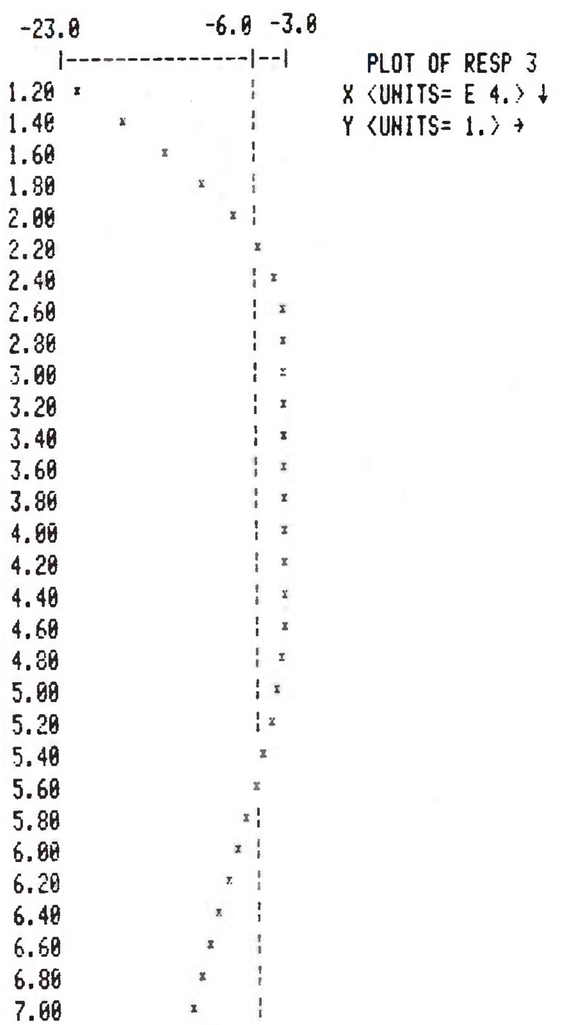


Figure 2. Frequency Response.

```
01♦LBL "CIRKT 3"   45 RCL 08
02 RCL 02          46 RCL 16
03 RCL 01          47 +
04 -               48 /
05 STO 03          49 STO 09
06 RCL 17          50 PI
07 *               51 /
08 RCL 01          52 RCL 03
09 RCL 02          53 /
10 *               54 "L1= "
11 SQRT            55 ARCL X
12 -               56 AVIEW
13 RCL 16          57 STO 14
14 *               58 STOP
15 2               59 RCL 09
16 /               60 RCL 03
17 PI              61 *
18 /               62 2
19 RCL 17          63 *
20 /               64 PI
21 RCL 01          65 *
22 /               66 1/X
23 RCL 02          67 "C2= "
24 /               68 ARCL X
25 "L2= "          69 AVIEW
26 ARCL X          70 STO 13
27 AVIEW           71 STOP
26 STO 12          72 RCL 03
29 STOP            73 4
30 2               74 /
31 *               75 PI
32 PI              76 /
33 *               77 RCL 01
34 RCL 02          78 /
35 RCL 01          79 RCL 02
36 +               80 /
37 2               81 RCL 09
38 /               82 /
39 *               83 "C1= "
40 RCL 17          84 ARCL X
41 *               85 AVIEW
42 STO 08          86 STO 15
43 RCL 16          87 END
44 *
```

REGISTERS	
R01	f_1
R02	f_2
R16	R_O
R17	Q_L

SIZE 018

Table I.

```
01♦LBL "RESP 3"   46 1/X        91 XEQ 01
02 FS? 00         47 -          92 R-P
03 GTO 00         48 RCL 14     93 RCL 16
04 "F=? "         49 RCL 19     94 *
05 PROMPT         50 *          95 LOG
06 ARCL X         51 STO 25     96 20
07 AVIEW          52 +          97 *
08♦LBL 00         53 RCL 24     98 FS? 00
09 2              54 RCL 25     99 RTN
10 *              55 RCL 17     100 "RESP= "
11 PI             56 /          101 ARCL X
12 *              57 +          102 "├ DB"
13 STO 19         58 STO 26     103 AVIEW
14 RCL 13         59 X<>Y       104 ADV
15 *              60 STO 27     105 RTN
16 CHS            61 X<>Y       106♦LBL 01
17 RCL 19         62 R-P        107 R-P
18 RCL 12         63 1/X        108 1/X
19 *              64 P-R        109 X<>Y
20 STO 21         65 1/X        110 CHS
21 1/X            66 X<>Y       111♦LBL 02
22 +              67 RCL 20     112 RDN
23 STO 20         68 +          113 RDN
24 1/X            69 1/X        114 R-P
25 RCL 16         70 X<>Y       115 R↑
26 1/X            71 1/X        116 *
27 RCL 21         72 RCL 21     117 RDN
28 RCL 17         73 XEQ 03     118 +
29 *              74 STO 24     119 R↑
30 STO 21         75 X<>Y       120 P-R
31 XEQ 03         76 STO 25     121 RTN
32 STO 24         77 X<>Y       122♦LBL 03
33 X<>Y           78 RCL 27     123 1/X
34 STO 25         79 RCL 26     124 +
35 RCL 16         80 XEQ 01     125 1/X
36 /              81 RCL 23     126 X<>Y
37 STO 23         82 RCL 22     127 1/X
38 X<>Y           83 R-P        128 X<>Y
39 RCL 16         84 X<>Y       129 1/X
40 /              85 XEQ 02     130 R-P
41 STO 22         86 RCL 24     131 1/X
42 RCL 25         87 RCL 18     132 P-R
43 RCL 15         88 +          133 RTN
44 RCL 19         89 RCL 25     134 .END.
45 *              90 X<>Y
```

REGISTERS	
R12	L_2
R13	C_2
R14	L_1
R15	C_1
R16	R_O
R17	Q_L
R18	R_L

SIZE 026

Note: Set flag 00 for PRPLOT

Table II.

Portable DC Magnetometer

The Electro-Mechanics Company is introducing their 6701 DC Magnetometer to fulfill a renewed interest in the EMI/RFI industry for a convenient and versatile magnetic field testing instrument.

Battery powered with rechargeable batteries, the 6701 is lightweight (7 lbs.) enough to be used in the field. It also performs with such high precision and reliability that the 6701 is ideal for the test lab.

The Application

Increasing applications of microelectronic technology is compounding the problems in electromagnetic interference (EMI). A great many of these difficulties are caused entirely by magnetic field radiation. The interference problems are compounding due to several factors: the increasing use of semiconductor devices in power switching applications, the trend toward miniature high density circuitry, high susceptibility of solid state components, and the general increase in sensitivity of new electronic systems. Many of the interfering fields are of an impulse nature and therefore contain a wide band of frequencies. Other problems may originate from low frequency magnetic phenomena by either inducing a current for circuit perturbation or by Hall effect perturbations within semiconductor junctions. A solution to these problems involves identifying the levels and sources of the interfering magnetic fields. This requires the use of an instrument which can measure a broad frequency range of low level magnetic fields.

Magnetometers are normally thought of as being very low frequency devices whose greatest use is to measure small variations in the earth's magnetic field. Although several types of magnetometers are available, rarely have they been developed for directional broadband low intensity pulse measurements and consequently, most instruments are not suitable for this type measurement. Many of the available instruments are incapable of being extended to broad frequency usage due to both theoretical and practical limitations.

The Instrument

The variable-mu magnetometer developed by the Electro-Mechanics Co. is based upon a principle significantly different from other magnetometers. Many of the desirable characteristics for a new approach to low frequency EMI are exhibited by the variable-mu technique.

The variable-mu magnetic sensor and receiver system called a Magnetic Field Intensity Meter or by its acronym MFIM is an instrument capable of measuring time varying magnetic fields from 1 Hz to 50 Khz. The sensor portion of the instrument is reasonably compact so that it may be placed close to small electrical components. The characteristics of being directionally sensitive (a cosine pattern about the sensor's axis) plus its compactness makes the instrument quite useful for identifying a source of a magnetic field.

Unlike most magnetic field sensor systems which can measure directional, time varying fields, the MFIM has a flat output independent of frequency. The instrument is a broadband receiver whose output is flat from the low to high frequency cut-off points. Broadband sensitivity of the instrument is approximately 300 picoteslas. Narrowband operation can be accomplished through the use of external filtering at the analog output of the receiver. Such filtering makes possible narrowband sensitivities of less than 100 picoteslas and with a dynamic range greater than 80 dB.

EMI Effects

Practically any electronic circuit subjected to time varying magnetic fields will have its operation influenced by the field. The influence which causes interference of the normal operation of the circuit may vary from a slight modulation of the output of the circuit to a complete degradation of the circuit operation. In the case of a very strong field, large voltages and currents may be induced into the circuitry and cause damage or destruction of the more sensitive components in the circuit. Current surges can cause thermal damage at points of high thermal inertia (poor heat dissipative characteristics), but the current path must have low resistance for the development of this type of damage. The greatest danger likely to affect a solid state device is a semiconductor junction puncture produced from induced voltages which exceed the peak voltage rating of the junction. A damaging high voltage induction is more often produced by a high intensity magnetic field, but it can also be generated by relatively low level fields which change intensity rapidly.

Consideration has been given tc damage and destruction of components brought about by high intensity magnetic fields or fields of rapidly changing levels; however, low level audio frequency fields are often all that is needed to disrupt proper operation of a solid state electronic circuit. Errant triggering of a digital microcircuit may cause a computer to malfunction or cause failure of a microcircuit control system. Failures caused by low level magnetic signals are, by far, the most prevalent and most overlooked problem sources in high density microelectronics. The proper architecture of component layout becomes as much of an art as an engineering task in high density electronics. The methods of reducing magnetic coupling, a spatial configuration, or the proper use of shielding materials can become as important as the intricacies of circuit design and function.

Intergated circuits can be affected by magnetic fields in several ways. Most IC semiconductors contain several types of components which are influenced by magnetic fields. These include Hall effect properties and magnetoresistive properties although the most prominent mode is loop coupling as developed in Faraday's law.

Although the level of induced voltage is proportional to the frequency of the interfering signal, it is fortunate that with nearly any shielding material, the shielding effectiveness of the shield increases with the increase in frequency.

Magnetic Field Measurement

It can be seen that miniature, high density electronic circuitry is susceptible to failure of operation or damage through the action of exposure to magnetic fields. The source of these fields needs to be found and measured in order that engineers can avoid problems in design before the problem surfaces.

The variable-mu magnetic field intensity meter is an instrument, developed by the Electro-Mechanics Co., which can detect levels and direction of the source of an interfering magnetic field in the ELF band from 1 Hz to 50 KHz.

Contact Mike Hart, Electro-Mechanics Company, P.O. Box 1546, Austin, Texas 78767, (512) 835-4684, or INFO/CARD #140.

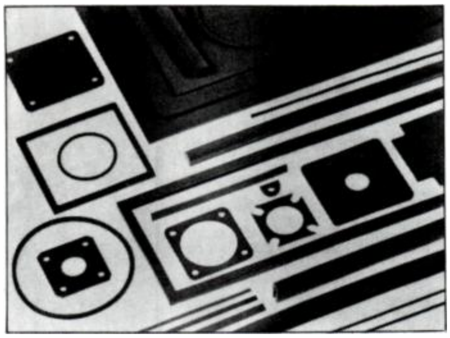
EMI Broadband Filters

U.S. Capacitor Corporation has introduced a comprehensive line of subminiature broadband EMI filters which are designed to meet or exceed applicable sections of MIL-F-28861 and MIL-F-15733 for high reliability.

Available in both solder-in and screw body filter styles, the line includes the most commonly used sizes and configurations and introduces a new high temperature series. **Contact Donald F. Hosmer, 11144 Penrose Street, Sun Valley, CA 91352. (213) 767-6770, or circle INFO/CARD #139.**

Silicone Elastomer For EMI Shielding

TECKNIT, recently announced the addition of a new silicone elastomer to its EMI Shielding product line. The new product, CONSIL-C, is a conductive elastomer filled with silver-plated copper particles, which achieves high-performance electrical conductivity and provides high-broadband shielding effectiveness. The product additionally provides moisture sealing for enclosure joints and seams. TECKNIT will manufacture its new CONSIL-C in the forms of sheets and die-cut flat gaskets as well as standard rectangular, round and cross-sectional strips. Strips and strip formed gaskets can also be provided with vulcanized joints to provide effectively continuous strip



lengths and finished gasket shapes. **Contact Peter Grant, EMI Shielding Products Manager, TECKNIT, 129 Dermody Street, Cranford, NJ, (201) 272-5500, or circle INFO/CARD #138.**

EMI Testing Service

Stanford Applied Engineering announces that full EMI testing services are now available to equipment manufacturers. The new testing laboratory offers certification testing to FCC Docket 20780 regulations, thus assuring the equipment manufacturer of complete FCC compliance. Documented FCC compliance is guaranteed. **Stanford Applied Engineering, 3520 De La Cruz Blvd., Santa Clara, CA 95050, (408) 988-0700, or please circle INFO/CARD #135.**

EMI And RFI Enclosures

EMI and RFI Shielding to satisfy new FCC and MIL-STD requirements is available, as a standard option, on plastic electronics enclosures from PAC-TEC Corp. Subsidiary of LaFrance Corp. Anticipating increased regulation of EMI and RFI, particularly in light of the economies offered by Plastics over metals in enclosure costs, PAC-TEC has thoroughly studied various shielding methods and now employs conductive coating applied to enclosure interiors. The coatings include applications of nickel, silver and graphite in various thicknesses. **PAC-TEC Corp., Enterprise and Executive Avenues, Philadelphia, PA 19153 U.S.A.: (215) 365-8400, or circle INFO/CARD #137.**

Conductive Epoxies For EMI Shielding

Key Polymer Corporation, has announced that its Marpoxy™ copper-filled conductive epoxy resins have passed Military Standard MIL-STD-461B for EMI (electromagnetic interference) shielding. The epoxies offer excellent shielding effectiveness over the frequency range from 150 KHz to 100 GHz, with attenuations from 65 dB to 76 dB. **Key Polymer Corp., Jacob's Way, Lawrence Industrial Park, Lawrence, MA 01842, (617) 683-9411, or circle INFO/CARD #134.**

Thermally Conductive Copper Filled Adhesive Film

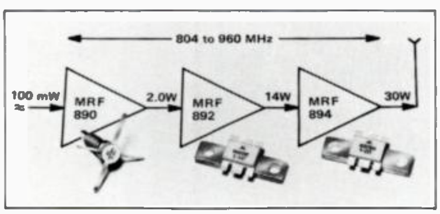
A glass supported, copper filled epoxy adhesive film has been developed as a less expensive alternative to silver filled adhesive systems by Ablestik Laboratories. Ablefilm® ECF550C provides the highest thermal conductivity of all the ABLESTIK adhesive films, making it ideal for substrate attach and heat sink bonding. In shielding applications, ECF550C provides electrical conductivity equal to silver at a significant cost savings. **Ablestik Laboratories, 833 West 182 Street, Gardena, CA, 90248 or circle INFO/CARD #136.**

TEM Cell

Two models of transverse-electromagnetic-mode (TEM) cells are now available from Amplifier Research. Models TC0250 and TC0500 are designed for rf testing of smaller objects within the frequency ranges of dc-250 MHz and dc-500 MHz respectively. Price for the TC0250 is $6600. For the TC0500, $5500. Delivery: 90 days ARO. **Amplifier Research, 160 School House Road, Souderton, PA 18694. 215-723-8181, or circle INFO/CARD #133.**

High Power 900 MHz Transistors

Motorola has introduced a new 24 Vdc, 900 MHz power transistor series. The new line includes the MRF 890, a 2.0 watt, 9.0 dB minimum gain **pre-driver,** the MRF 892, a 14 watt, 8.5 dB **driver,** and the MRF 894, a 30 watt, 7.0 dB **final amplifier.** The new RF transistors are fully characterized across



the 804 to 960 MHz frequency range and are intended for large-signal, common-base amplifier applications in industrial and commercial cellular FM radio-telephone equipment, which was recently approved by the FCC. All devices have guaranteed gain performance at 900 MHz, collector efficiencies of 55% minimum, and will withstand 30:1 VSWR load mismatch at rated output power and voltage supply. Sampling is now taking place with approximately 8 to 12 weeks delivery. Prices in 100-499 pc. qty., MRF 890-$10.00, MRF 892-$19.55, MRF 894-$30.60. **Contact Tom Bishop, Motorola Semiconductor Products Inc., P.O. Box 20912, Phoenix, Arizona 85036, (602) 244-6394 or circle INFO/CARD #132.**

Miniature RF Coaxial Connector Kit

E.F. Johnson Company, Waseca, MN announced the introduction of its

(Continued on page 46.)

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Reader Contest Rules

- Examine this issue of *r.f. design* carefully. Pick the ten advertisements you think your fellow engineers-subscribers will best remember.
- List them on the entry form in this issue. Every entry form must be filled in completely or it will not be considered.
- To enter, readers must be engaged in or supervising design activities or involved in component specification/selection for design activities.
- Contest void where prohibited or taxed by law. Liability for any taxes is the sole responsibility of the winner.
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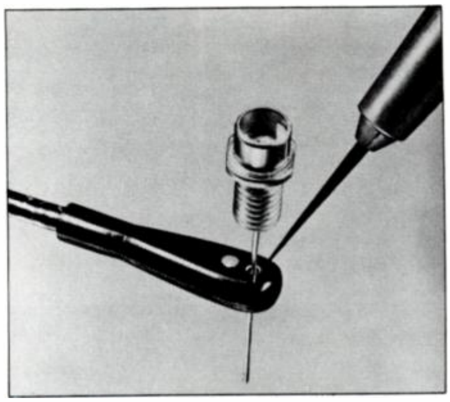
INFO/CARD 23

CODE-A-PHONE®

(Continued from page 41.)
JCM Miniature RF Coaxial Kit. The kit contains 5 pieces each of 9 popular styles of Gold Plated Miniature RF Coaxial Connectors. All 45 pieces are housed in a handy 9 drawer plastic cabinet. Complete assembly instructions are enclosed in each drawer. All connectors are 50 ohm type and the Plug and Jack assortment covers the major cable group sizes—RG-178, RG 196, RG-174, RG-188, RG-316, RG-55, RG-58, RG-141, RG-142, RG-233. The complete kit is OEM priced at $98.40 and is available through the nearest Johnson Distributor. **E.F. Johnson Company, 299 10th Ave., S.W., Waseca, MN 56093, (507) 835-6222, or circle INFO/CARD #131.**

Wideband Current Probe

The Model 711 Miniature Wideband Current Probe measures wideband current or pulses without loading the circuit being tested. It induces no appreciable capacitive nor inductive effects on circuitry, therefore the signal under measurement does not change. Being 20 times smaller than other commercially available devices, it requires no additional lead length for insertion and, therefore it does not alter circuit performance. The 711



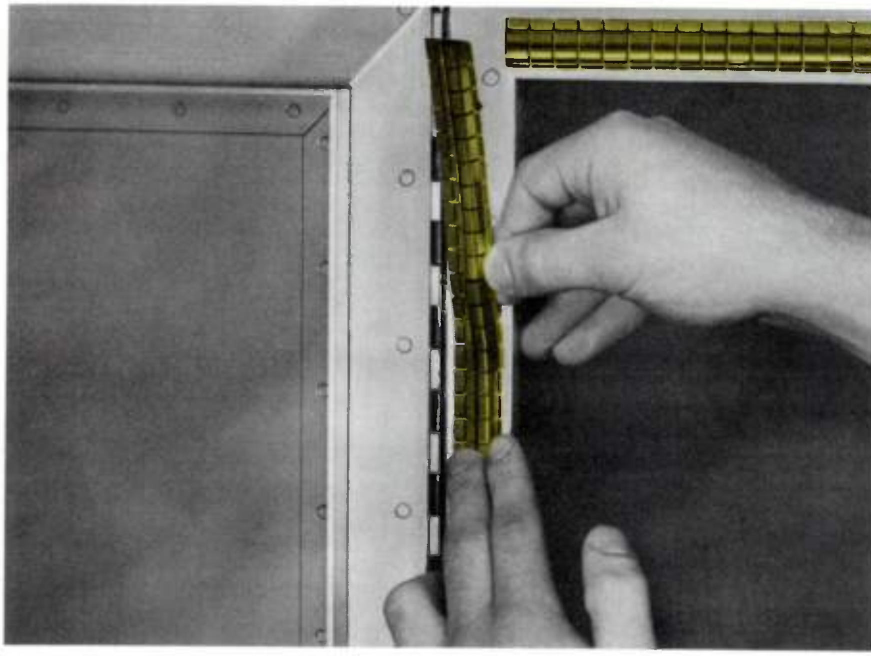
exhibits only .02Ω shunted by 4μh insertion impedence. The current is sensed by placing the conductor through the center of the probe, which will accept a maximum lead diameter of #20 AWG. Typical applications include: current measurement in high frequency amplifiers, drive circuits for LEDs and lasers, electro-magnetic devices (deflection coils), energy storage devices, and SCR gate drivers. The 711 is used in any field requiring the accurate measurement of current up to 100 amps with Ampere-Sec. products of less than 6A-μs. **American Laser Systems Inc., 106 James Fowler Rd., Santa Barbara Airport, Goleta, CA 93117, (805) 967-0423 or circle INFO/CARD #130.**

S Parameter Design Program

Gamma Electronics Inc., announces that it will market several of its' design and analysis computer programs. The first program to be offered is called SPAMO (S Parameter Analysis and Manual Optimization). The program runs on a low cost computer requiring only 32 K of memory and a single disk drive. The program contains several unique features: direct or computed S parameter data entry, easy topology entry (R/L/C, node #, node #, value), add additional elements and nodes at any time as well as changing component types and values, execution time of approximately 20 seconds per frequency. The manual that is supplied with the program provides several step by step examples. One example starts with a "bare" transistor then adds and modifies components to obtain a final design which has a gain of 11 dB minimum from 50 to 1000 Mhz, a flatness of ± 0.05 dB and an input/output VSWR of 1.6:1 maximum. **Gamma Electronics Inc., 6223 Edgewater Dr., Falls Church, VA 22041, (703) 820-7741 or INFO/CARD #129.**

Solid State Dip Meter

A solid state dip meter for testing radio frequencies, antennas, oscil-



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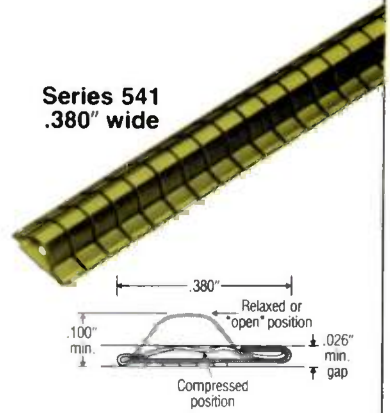

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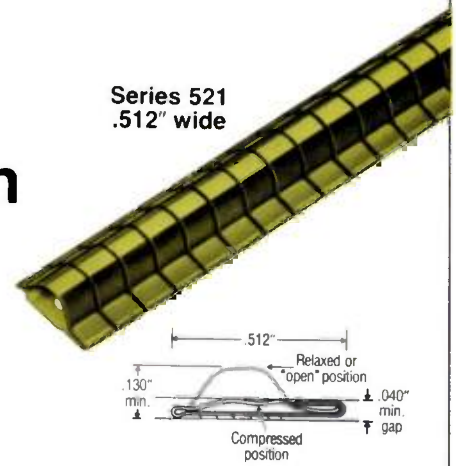
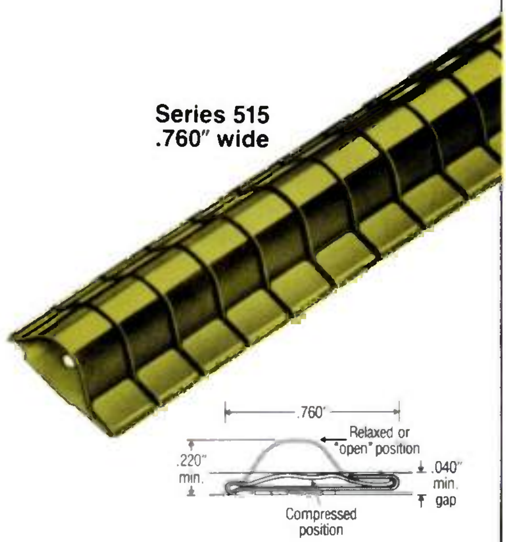


Series 515
.760" wide



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INFO/CARD 26

lators, RF chokes, and etc., formerly manufactured by Millen, is available from Caywood Electronics, Inc. The Millen Solid State Dipper is a portable oscillating frequency meter that determines the resonant frequency of de-energized resonant circuits with an accuracy of ±2%. Covering a range from 1.65 to 310 MHz with 7 plug-in coils, it also features an absorption-type wavemeter with the oscillator circuit acting as a Q multiplier amplifier to enhance tuning response and dip sensitivity. The Millen Solid State Dipper is priced at $210 each, complete with 7 coils and case. **Contact Wade Caywood Electronics, Inc., P.O. Drawer U, Malden, MA 02148-0921, (617) 322-4455, or INFO/CARD #127.**

SAW Filter

RF Monolithics, Inc., has announced a new clean-up filter to eliminate the adjacent channel interference which results from certain combination of TV receivers and cable converters. Called the CTVF-3, the filter is fully matched, SAW stabilized and carries an insertion loss of less than 5 dB. It is priced at $16.25 in quantity. **RF Monolithics, Inc., 4441 Sigma Rd., Dallas, Texas 75234, (214) 233-2903 or circle INFO/CARD #126.**

New Literature

EMI Consultant Service Brochure

R & B Enterprises, long a leader in EMI testing, consulting, publications, and related training, announces a new brochure which details its comprehensive services. **R & B Enterprises, 1050 Colwell Lane, Conshohocken, PA 19428. (215) 828-6236 or circle INFO/CARD #107.**

EMI Shielding Design Guide

TECKNIT, announces the availability of a complete reference source that provides design engineers with up-to-date, state-of-the-art information for dealing with the increasing demands of EMI shielding problems. The 75-page TECKNIT publication systematically addresses the design criteria involved in the area of EMI shielding. A detailed tutorial reference for those unfamiliar with the selection and application of EMI shielding materials, the guide is organized into ten (10) convenient sections. Each section covers a specific topic in sufficient detail to provide readers with information on subjects ranging from "Metal Shielding Barriers" to "Shielding Performance Testing" and "Procurement." The TECKNIT Design Guide is available for $10.00. To obtain your copy, **contact Robert E. Bilby, Manager Corporate Communications, TECKNIT, 129 Dermody Street, Cranford, NJ (201) 272-5500.**

55-85 MHz 75 Ohm Amplifier Application Note

California Eastern Laboratories, Inc., offers new Application Note AN82302 titled "A 55-85 MHz 75 OHM AMPLIFIER USING TWO NE74114 BIPOLAR MICROWAVE TRANSISTORS." This 70 MHz IF amplifier can be used in many different systems applications, including a TVRO front end. **California Eastern Laboratories, Inc., 3005 Democracy Way, Santa Clara, CA, 95050, (408) 988-3500, or circle INFO/CARD #105.**

Microwave Transistor Selection Guide

A new microwave transistor selection guide is now available. **Motorola Semiconductor Products Inc., P.O. Box**

20912, Phoenix, AZ 85036, (602) 244-6900, or circle INFO/CARD #104.

Automatic Power Meter Calibrator Brochure

A new 12 page brochure describing the unique System II Automatic Power Meter Calibrator has just been released by Weinschel Engineering. It demonstrates how System II can complete power meter calibration tasks in minutes providing accurate, cost effective calibration of terminating power meters (even by semi-skilled personnel) at a rate of 1.5 seconds per point with typical accuracies of less than 1% above those available directly from the National Bureau of Standards including mismatch uncertainties. **Weinschel Engineering, One (1) Weinschel Lane, Gaithersburg, MD 20877. (301) 948-3434 or INFO/CARD #106.**

Eimac Cavity Amplifiers

Eimac, a division of Varian Associates, describes its complete line of cavity amplifiers designed for FM, TV and UHF broadcast service in a new 14-page brochure. **Varian Associates, EIMAC Division, 301 Industrial Way, San Carlos, CA 94070, (415) 592-1221, or circle INFO/CARD #102.**

Instrument Catalog

The latest full-line instrumentation catalog from Marconi is now available. This 176 page book is of particular interest to signal generator users, as it includes complete details of both the new cavity tuned 2017 generator and the 2018/2019 synthesizers. **Marconi Instruments, 100 Stonehurst Court, Northvale, N.J. 07647, (201) 767-7250, or INFO/CARD #101.** □

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