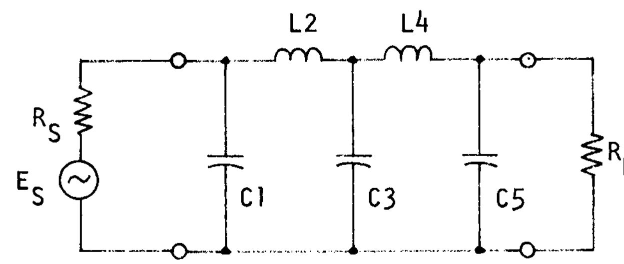
A COMPUTERIZED LOW-PASS FILTER DESIGN PROCEDURE

INTRODUCTION

Since the 1963 publications of Geffe's *Simplified Modern Filter Design*[1] and White's *Handbook on Electrical Filters*[2], the design of passive L-C filters has been relatively simple and straightforward for the average electronics engineer. However, the realization of an optimum design, that is, a design requiring a minimum number of standard value capacitors, continues to remain a problem. An example of an optimum filter design is shown in Figure 1. Here the three capacitance values can each be realized with single, standard-value, commercially available capacitors. Of course, any required filter capacitance can be realized by paralleling capacitors, but this technique has several disadvantages. In addition to the greater expense and space required by the extra capacitors, the parallel combination becomes resonant at some frequency. In the case of the low-pass filter, the resonance occurs somewhere in the filter stopband, thus creating a high shunt impedance (when it should be very low) with a corresponding undesired hole in the filter stopband. Obviously, using single capacitors of correct value to provide the required filter shunt elements is preferable to using paralleled capacitors. But, how can this be easily accomplished? This article will discuss how a computer can be programmed to select filters requiring only standard capacitor values for any specified value of termination resistance.

The five-element, capacitive-input and output, low-pass filter will be used to demonstrate how a computer can be made to select filters that require only standard value capacitors. This particular filter type is easy to construct and is not too sensitive to component value variation, while still providing adequate signal selectivity for most non-stringent filtering applications. Usually, it will be satisfactory if the attenuation roll-off is at least 30 dB/octave and if the actual cutoff frequency is within ±5% of the desired cutoff frequency. The 3-capacitor and 2-inductor filter configuration is generally preferable to its dual configuration (3-inductor and 2-capacitor), because minimizing the number of inductors minimizes the component cost and improves the filter performance (inductors are more costly and have greater loss than capacitors).



C1, C5 = .068μF

C3 = .22 μF

L2, L4 = .445mH

$R_S = R_L$ = 50 ohms

$f_{co_{3dB}}$ = 28.9 kHz

C3/C1 = 3.24

Figure 1. An Optimum Filter Design (only standard-value capacitors required)

FIVE-ELEMENT CHEBYSHEV LOW-PASS FILTER

The 5-element Chebyshev low-pass filter will be used in this design procedure. The attenuation response of this filter type is characterized by a specific level of maximum passband attenuation ripple (A_{peak} or A_p) of usually one dB or less, and a constantly increasing attenuation above the cutoff frequency. For this filter type, the cutoff frequency is defined as that frequency where the attenuation first exceeds the peak attenuation level of the passband ripple, A_p. There is a particular condition that exists when the filter passband attenuation has maximum flatness, or a passband ripple of zero dB. The filter having this particular characteristic is known as a Butterworth filter, and, by common agreement, the cutoff frequency is specified to occur at the 3 dB attenuation level. The equations for the Butterworth and Chebyshev filter element values have been solved and the results published in catalogs[1,2,3,4] which have been available to the filter designer for many years. The published element values (designated as G1, G2, G3, G4, and G5 in the catalogs) have been normalized for ω_{co}= one radian/second and $R_S=R_L$= one ohm. There is, of course, only one catalog of normalized element values for the 5-element Butterworth filter, but there can be an infinite number of catalogs for the Chebyshev filter since there can be an infinite number of different levels of maximum passband attenuation ripple. Usually, only eleven of the Chebyshev normalized catalogs are published, and these particular catalogs are associated with the reflection coefficient, VSWR, and A_p values listed in Table 1. Some other publications give the normalized values for A_p values of .1, .25, .5, and 1 dB and higher where the level of maximum passband attenuation ripple may be of primary interest. The reflection coefficient, VSWR, and A_p values are all inter-related and are associated with a particular Chebyshev catalog. Thus, either of these three parameters may be used to specify a particular catalog. In this article, the reflection coefficient (in percent) will be used as the governing parameter, since it is used as such in the design equations to follow. As seen from Table 1, there are only eleven distinct Chebyshev catalogs that are commonly published (not counting the Butterworth), but it is obvious that there can be many more. To obtain the unpublished catalogs, the general design equations for the Chebyshev filter must be used. The dependency on the published data will then be removed and it will be possible to make a selection from an infinite number of catalogs to exactly match a particular design requirement.

Table 1. A listing of commonly published Chebyshev R.C., VSWR and A_p values.

R.C. (%)	VSWR	A_p (dB)
1	1.020	.000434
2	1.041	.00174
3	1.062	.00391
4	1.083	.00694
5	1.105	.0108
8	1.174	.0277
10	1.222	.0436
15	1.353	.0988
20	1.500	.177
25	1.677	.280
50	3.000	1.249

COMPUTER PROGRAM FOR CHEBYSHEV NORMALIZED FILTER VALUES

The Chebyshev element value design equations given in Seymour Cohn's paper, *Direct Coupled Resonator Filters*[5], were solved for an equally loaded filter of five elements in terms of the reflection coefficient. A BASIC computer program was written using these equations to solve and tabulate the VSWR, peak ripple, normalized component values, and the ratio of the 3 dB cutoff frequency to the A_p cutoff frequency for any specified value of reflection coefficient. As an additional design reference aid, the ratio of the normalized values of C3 and C1 was included in the tabulation. The computer program consists of 19 statements, and is listed with its print-out of normalized component values in Table 2. Values of VSWR, 3 dB to A_p cutoff frequency ratio, A_p ripple, normalized filter element values (G1 to G5) and the G3/G1 ratio are tabulated for those values of reflection coefficient (R.C.)

BASIC Program #1 for calculation of normalized filter element values

```
10 PRINT" VSWR    R.C.   A3DB/AP RIPPLE  NORMALIZED C1&5, C3, L2&L4 "
20 PRINT" ----    (%)    ------   (DB)   (G1 & G5)  (G3)    (G2&G4)  G3/G1"
30 DATA .4809,1,1.518,2,2.5,3,3.5,4,4.79586,5,6,7,8,10,15,15.0875
40 DATA 18,20,25,25.836,28,30,32.9773,40,45.3513
50 PRINT " 1.000   .000  2.000   .00000   .3090   1.000     .8090  3.24"
60 READ E
70 V=(1+E/100)/(1-E/100)
80 A=-4.3429*LOG(1-(E/100)^2)
90 K=1.33/A^.1
100 D=.5*(K+1/K)
110 B=A/17.37
120 X=LOG((EXP(B)+EXP(-B))/(EXP(B)-EXP(-B)))
130 Y=.5*(EXP(.1*X)-EXP(-.1*X))
140 G1=.61804/Y
150 G2=(4*.30902*.80902)/((Y*Y+.34549)*G1)
160 G3=(4*.80902)/((Y*Y+.90451)*G2)
170 PRINT, 180, V,E,D,A,G1,G3,G2,G3/G1
180 FMT F6.3,F7.3,X1,F6.3,X1,F8.5,X1,F7.4,X2,F7.4,X2,F7.4,X1,F5.2
190 GO TO 60
```

BASIC Program #1 Print-out of normalized filter element values

VSWR	R.C.	A3DB/AP	RIPPLE	NORMALIZED C1&5,	C3,	L2&L4	
----	(%)	------	(DB)	(G1 & G5)	(G3)	(G2&G4)	G3/G1
1.000	.000	2.000	.00000	.3090	1.000	.8090	3.24
1.010	.481	1.819	.00010	.4080	1.0895	.9283	2.67
1.020	1.000	1.616	.00043	.4871	1.2258	1.0499	2.52
1.031	1.518	1.515	.00100	.5428	1.3103	1.1220	2.41
1.041	2.000	1.455	.00174	.5848	1.3691	1.1693	2.34
1.051	2.500	1.409	.00272	.6225	1.4187	1.2068	2.28
1.062	3.000	1.374	.00391	.6560	1.4606	1.2362	2.23
1.073	3.500	1.345	.00532	.6866	1.4975	1.2602	2.18
1.083	4.000	1.322	.00695	.7148	1.5305	1.2799	2.14
1.101	4.796	1.291	.01000	.7564	1.5773	1.3049	2.09
1.105	5.000	1.284	.01087	.7665	1.5885	1.3104	2.07
1.128	6.000	1.256	.01566	.8132	1.6390	1.3322	2.02
1.151	7.000	1.233	.02133	.8566	1.6846	1.3480	1.97
1.174	8.000	1.214	.02788	.8973	1.7265	1.3593	1.92
1.222	10.000	1.184	.04365	.9732	1.8032	1.3723	1.85
1.353	15.000	1.136	.09883	1.1440	1.9722	1.3715	1.72
1.355	15.087	1.136	.10000	1.1469	1.9750	1.3712	1.72
1.439	18.000	1.117	.14304	1.2396	2.0667	1.3580	1.67
1.500	20.000	1.107	.17729	1.3019	2.1286	1.3456	1.63
1.667	25.000	1.086	.28028	1.4560	2.2834	1.3066	1.57
1.697	25.836	1.083	.30002	1.4817	2.3095	1.2992	1.56
1.778	28.000	1.077	.35457	1.5487	2.3780	1.2793	1.54
1.857	30.000	1.071	.40958	1.6112	2.4424	1.2599	1.52
1.984	32.977	1.063	.50000	1.7058	2.5409	1.2296	1.49
2.333	40.000	1.049	.75720	1.9404	2.7899	1.1531	1.44
2.660	45.351	1.041	1.00000	2.1350	3.0010	1.0911	1.41

Table 2. Program #1 and Print-out
BASIC Program #1 for calculation of normalized filter element values

given in data statements 30 and 40 of the program. A number of R.C. values in percent were selected to give a representative tabulation of filter values that could be checked for accuracy against previously published authoritative tables. For this reason, the R.C. values of 1, 2, 3, 4, 5, 10, 15, 20, and 25 were used to allow the tabulated values of G1, G3, and G2 to be checked against Saal's catalog to confirm that Program #1 does give valid results. By comparing the Program #1 calculated values with the corresponding values given in Saal's catalog, it can be demonstrated that the Program #1 calculated values are either identical or differ occasionally only in the last decimal place. Because the equations in statements 70-160 cannot be used to calculate the Butterworth catalog, this catalog (corresponding to R.C. = zero) was calculated separately and put in statement 50. The computer was programmed to print the Butterworth values at the head of the tabulation. Because the Butterworth tabulated value of A_{3dB}/A_p is 2.000, the values of G1-G5 are one-half of the values that are customarily published for the Butterworth normalized values. Values of R.C. between zero and .48 were not used because the calculated normalized values of G1-G5 become less correct as the R.C. approaches zero.

LOW-PASS FILTER DESIGN EXAMPLE

An example of a filter design using the tabulated normalized filter values is shown in Figure 2. Here the Butterworth normalized values are used to calculate the component values of a low-pass filter having a 3 dB cutoff of 28.9 kHz and resistive terminations of 50 ohms. The cutoff frequency of 28.9 kHz was selected (from the tabulation of Appendix 3 for the 50-ohm termination) to assure that the resulting capacitor values would turn out to be standard commercially available values. This makes the filter especially convenient to construct. Of course, if a cutoff frequency other than 28.9 kHz is desired, this particular design won't be of immediate interest. However, the many different Chebyshev filter catalogs can be searched for a combination of capacitors which will yield an optimum design having a cutoff frequency usually within ±5% of the desired cutoff frequency. Because these filters are intended for non-stringent applications, it generally will be acceptable if the actual cutoff frequency is

From Print-out of Program #1 —

GIVEN: $G1,5 = 0.309$

$G3 = 1.000$

$G2,4 = 0.809$

$f_{co_{3dB}}/f_{co_{A_p}} = 2.000$

LET: $R_S = R_L = 50$ ohms

$f_{co_{3dB}} = 28.9$ kHz

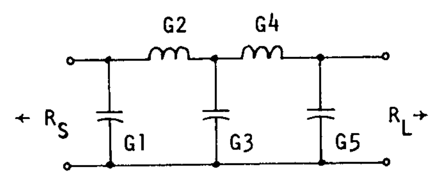
THEN: $f_{co_{A_p}} = 28.9k/2 = 14.45k$

$\omega = 2\pi f_{co_{A_p}} = 90.8k$

$R/\omega = 50/90.8k = (.551)10^{-3}$

$1/R\omega = 1/(50)(90.8k) = (.22)10^{-6}$

Normalized Filter



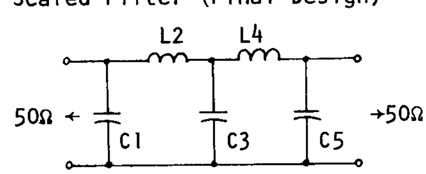
Filter Component Value Calculations

$L2,4 = G2(R/\omega) = .809(.551)10^{-3} = .445$mH

$C1,5 = G1(1/R\omega) = .309(.22)10^{-6} = .068\mu$F

$C3 = G3(1/R\omega) = 1.00(.22)10^{-6} = .22\mu$F

Scaled Filter (Final Design)



$f_{co_{3dB}} = 28.9$ kHz

Figure 2. An example of filter design calculations based on Program #1 print-out.

within 5% of the desired frequency. Also, the actual component values can differ by a few percent from the exact design values without causing any problems. For example, if a filter having a 3 dB cutoff frequency of 20 kHz is desired, it could be obtained with a Butterworth design using .10 and .33 uF capacitors. Although the requird C3/C1 ratio for a Butterworth design is 3.24 (see the tabulation of Program #1), the capacitor ratio of 3.30 is within 2% of the exact ratio and is close enough to use.

COMPUTER PROGRAM FOR OPTIMUM LOW-PASS FILTER DESIGNS

By using the program of Table 2, it is possible to obtain any desired number of Chebyshev normalized designs. The increments of the reflection coefficients (in data statements 30 and 40 of program #1) should be made small enough so that the resulting normalized component values increase in small steps. Some trial and error in the selection of the R.C. is advisable to assure that all possible designs will be considered. Because standard capacitors are available only in a limited number of values, only a limited number of filter designs are possible. The problem is to find out what these designs are for every R.C. and for the particular resistive termination required for the filter. This is where Computer Program #2 comes in.

Computer Program #2 (see Table 3) is used for calculating filter designs requiring only standard value capacitors for any specified filter termination. Table 4 shows the abbreviated print-out of the results of program #2 for two values of termination resistance - 50 and 500 ohms. This program calculates all the filter designs that are possible for: (1) the capacitor values listed in data statements 10 to 60, (2) the reflection coefficients listed in data statements 170 and 180, and (3) the resistance termination specified by the user in statement 130. The commercially available capacitor values are listed once in the data statement and remain unchanged. The R.C. values are also similarly listed. The only remaining input required is that of the termination resistance. This can be any number, but it must be specified and entered at the beginning of each computer run.

The program uses each value of R.C. to calculate all those parameters common to that particular value of R.C. These parameters and their tabular headings are then printed in the first two lines of the print-out. The computer next compares all possible ratios of the capacitor values in data statements 10 to 60 with the previously calculated G3/G1 ratio. When a match is found within ±2%, the values of C1, C3, L2 and C3/C1 are calculated and tabulated. Also calculated and tabulated are the A_p and 3 dB cutoff frequencies of the filter. If the specified termination resistance is greater than 300 ohms, the frequency and inductance headings are printed in kHz and millihenries. If the termination resistance is 300 ohms or less, the headings are printed in MHz and microhenries. After all possible combinations of capacitance within the ±2% window have been accounted for, the computer goes to the next R.C. value and the process is repeated.

BASIC Program #1 Print-out of normalized filter element values

```
 10  DATA .000,2.000,.3090,1.000,.8090,1.000,.000
 20  DATA .0010,.0011,.0012,.0013,.0015,.0016,.0018,.0020
 30  DATA .0022,.0024,.0025,.0027,.0030,.0033,.0036,.0039
 40  DATA .0043,.0047,.0050,.0051,.0056,.0062,.0068,.0075
 50  DATA .0082,.0091,.010,.012,.015,.018,.020,.022,.027
 60  DATA .033,.047,.050,.068,.10,.15,.22,.33,.47,.68,1.0
 70 DIM A(44)
 80 READ E,D,G1,G3,G2,V,A
 90 FOR X=1 TO 44
100 READ A(X)
110 NEXT X
120 PRINT "SPECIFY LOW-PASS FILTER TERMINATION RESISTANCE IN OHMS"
130 INPUT R8
140 S=1E-3
150 IF R8>300 THEN 170
160 S=1
170  DATA 0,.4809,1.518,4.79586,5,8,10,15,15.0874,25.8353,32.9773
180  DATA 45.3513
190 READ E
200 Z=0
210 IF E<.2 GO TO 320
220 V=(1+E/100)/(1-E/100)
230 A=-4.3429*LOG(1-(E/100)^2)
240 K=1.33/A^.1
250 D=.5*(K+1/K)
260 B=A/17.37
270 X=LOG((EXP(B)+EXP(-B))/(EXP(B)-EXP(-B)))
280 Y=.5*(EXP(.1*X)-EXP(-.1*X))
290 G1=.61804/Y
300 G2=(4*.30902*.80902)/((Y*Y+.34549)*G1)
310 G3=(4*.80902)/((Y*Y+.90451)*G2)
320 LET R =G3/G1
330 R1=.98*R
340 R2=1.02*R
350 FOR X=44,1,-1
360 FOR W=1,44
370 B=A(X)/A(W)
380 IF B<R1 THEN 570
390 IF B>R2 THEN 570
400 IF Z=1 THEN 520
410 Z=1
420 PRINT" R.C.(%)  A3DB/AP  G1&G5    G3      G2&G4    VSWR     AP(DB)";
430 PRINT "  G3/G1"
440 PRINT, 450, E,D,G1,G3,G2,V,A,G3/G1
450 FMT F7.3,X3,F5.2,X2,F7.4,X1,F7.4,X1,F7.4,X1,F6.3,X2,F7.4,X3,F5.2
460 PRINT " FCO(3DB) FCO(AP)    C1 & C5       C3          L2 & L4"
470 IF R8>300 THEN 510
480 PRINT "    (MHZ)    (MHZ)    (MICRO-F)  (MICRO-F)  (MICRO-H)   C3/C1"
490 GO TO 520
500 IF Z=1 THEN 520
510 PRINT"    (KHZ)    (KHZ)    (MICRO-F)  (MICRO-F)  (MILLI-H)   C3/C1"
520 LET F=(D*.1592*G1/R8)/A(W)/S
530 LET L=S*R8*R8*A(W)*G2/G1
540 PRINT,550, F,F/D,A(W),A(X),L,A(X)/A(W)
550 FMT F8.3,F8.3,F10.4,F10.4,F13.2,F9.2
CONTINUED ABOVE

CONTINUED FROM BELOW
560 Z=1
570 NEXT W
580 NEXT X
590 PRINT
600 GO TO 190
    **END**
```

Table 3. BASIC Program #2 for calculation of filters requiring standard-value capacitors

```
SPECIFY LOW-PASS FILTER TERMINATION RESISTANCE IN OHMS
!50
 R.C.(%)   A3DB/AP   G1&G5      G3        G2&G4    VSWR      AP(DB)    G3/G1
  .000      2.00     .3090   1.0000      .8090   1.000     .0000     3.24
 FCO(3DB)  FCO(AP)    C1 & C5         C3         L2 & L4
  (MHZ)     (MHZ)    (MICRO-F)    (MICRO-F)    (MICRO-H)     C3/C1
   .020      .010      .1000        .3300        654.53       3.30
   .029      .014      .0680        .2200        445.08       3.24
   .042      .021      .0470        .1500        307.63       3.19
   .197      .098      .0100        .0330         65.45       3.30
   .240      .120      .0082        .0270         53.67       3.29
   .289      .145      .0068        .0220         44.51       3.24
   .317      .159      .0062        .0200         40.58       3.23
   .351      .176      .0056        .0180         36.65       3.21
   .419      .209      .0047        .0150         30.76       3.19
   .787      .394      .0025        .0082         16.36       3.28
  1.230      .615      .0016        .0051         10.47       3.19
  1.640      .820      .0012        .0039          7.85       3.25
  1.789      .894      .0011        .0036          7.20       3.27
  1.968      .984      .0010        .0033          6.55       3.30

 R.C.(%)   A3DB/AP   G1&G5      G3        G2&G4    VSWR      AP(DB)    G3/G1
  .481      1.82     .4080   1.0895      .9283   1.010     .0001     2.67
 FCO(3DB)  FCO(AP)    C1 & C5         C3         L2 & L4
  (MHZ)     (MHZ)    (MICRO-F)    (MICRO-F)    (MICRO-H)     C3/C1
   .236      .130      .0100        .0270         56.89       2.70
   .288      .158      .0082        .0220         46.65       2.68
   .315      .173      .0075        .0200         42.67       2.67
   .348      .191      .0068        .0180         38.68       2.65
   .422      .232      .0056        .0150         31.86       2.68
   .945      .520      .0025        .0068         14.22       2.72
  1.477      .812      .0016        .0043          9.10       2.69
  2.363     1.299      .0010        .0027          5.69       2.70


SPECIFY LOW-PASS FILTER TERMINATION RESISTANCE IN OHMS
!500
 R.C.(%)   A3DB/AP   G1&G5      G3        G2&G4    VSWR      AP(DB)    G3/G1
  .000      2.00     .3090   1.0000      .8090   1.000     .0000     3.24
 FCO(3DB)  FCO(AP)    C1 & C5         C3         L2 & L4
  (KHZ)     (KHZ)    (MICRO-F)    (MICRO-F)    (MILLI-H)     C3/C1
  1.968      .984      .1000        .3300         65.45       3.30
  2.894     1.447      .0680        .2200         44.51       3.24
  4.187     2.093      .0470        .1500         30.76       3.19
 19.677     9.839      .0100        .0330          6.55       3.30
 23.997    11.998      .0082        .0270          5.37       3.29
 28.937    14.468      .0068        .0220          4.45       3.24
 31.737    15.869      .0062        .0200          4.06       3.23
 35.138    17.569      .0056        .0180          3.67       3.21
 41.866    20.933      .0047        .0150          3.08       3.19
```

Table 4. Abbreviated Print-out of Program #2 for 50 and 500-ohm resistive terminations

The resulting list of valid filter designs can be rather long, but the user can quickly scan the extreme left-hand column of f_{co}(3 dB) to find the filter which most closely approaches the desired cutoff frequency. The tedious and error-prone filter calculations are thus eliminated and, as an additional bonus, only standard value capacitors are needed.

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PROGRESS IN EMC FILTERING

Introduction

Conventional frequency-selective filtering, based on lumped L's and C's, was developed about 50 years ago for communications. It had a solid mathematical foundation and was based expressly on impedance match which the communications engineer could control. In the meantime, filtering for communications has made great strides in sophistication; e.g., correlation techniques and praetersonics.

In contrast, some of the supposedly simple filtering for EMC power feedlines, permeating the whole system, is still in the state it was half a century ago. It is a severe mistake - and a matter of severe practical consequences - to equate filtering for communications with filtering for EMC for which, by its very nature, impedance matching does not hold. In power feed lines, the EMC engineer cannot enforce impedance match nor does he often know what the interface impedances are at any given time. Yet, there are EMC books that discuss EMC filtering in terms of classical filter theory. Such simplifying assumptions are as wrong as they are convenient. Still, today, design and testing of interference filters is done according to military standards based on meaningless, unrealistic, and uncritical conditions. To "compensate" for the uncertainty of the boundary conditions and their "unpredictable" effects (errors of ± 50 dB or more), some EMC engineers build much bigger filters than necessary or test the filters with incommensurably great accuracy, in neither case achieving what they intend to achieve.

In some circles, EMC is looked upon with disdain - and with some justification - because inappropriate standards permeate not only filtering, as we shall see in all its defectiveness, but also grounding, bonding, and shielding (3). Many standards are in need of updating. But more is involved: EMC is a complex, multifaceted affair of interacting objectives, often contrary to the primary mission of the system under consideration. All the more, it is necessary to realize what standards are: Rules made for simplifying assumptions (which may create other problems by solving the one on hand) and by committees (where concensus often means compromise with less competent, but vocal members). A good EMC engineer, as it is true in all professions, works systemicly from sound principles. An average or bad EMC engineer goes "by the book" by relying on often unavoidably imperfect standards that cannot anticipate all ramifications enountered in a particular system. Adaptive thinking in sound principles, and making the right simplifications are what counts in EMC, not blind reliance on cookbook routines. (After this brief philosophical discourse, let us go back to the specific problem on hand.)

Many attempts have been made to solve the problem of EMC filter testing, but the solutions were either too simplified to be meaningful or too general to be economical. The consequences of inappropriate filter testing are: One never knows whether the filter works in the stopband or does it ring (or have insertion gain) in the pass band such that the filter created new, unsuspected interference or breaks down. In many an installation, filters had to be thrown out (TEMPEST in particular) because they caused more harm than good, although they were successfully tested according to military standards.

The proper approach to filtering is to:

(1) Throw away conventional filter books and erroneous filter standards.
(2) Clearly delineate the ranges of interface impedances (source and load impedances).
(3) Establish a sufficiently broad statistical data base, but do not fall into the trap of calculating the mean or average values.
(4) Develop a transparent worst-case theory with boundary conditions according to (2) & (3).
(5) Based on (4) set up simple and predictive guidelines for cost-effective design and practicable testing of interference filt

Data base, theory and design guidelines

Figures 1 and 2 indicate the ranges of normal mode and common mode interfacial impedances for 60 Hz, 120 V sources and loads for the critical frequency range of 1 to 100 kHz. They are extracted from a broad statistical data base from measurements in factories, laboratories, households, and aboard ships. (More details are provided in reference 3.) It is obvious that filters do not meet with ANY impedance, rather the ranges of real impedances are quite delineated. Another crucial point is not conveyed in these figures; namely (except for common mode loads), the Q's of the interface impedances are mostly much below 2, occasionally some higher one, but not very much so.

The values (as, function of frequency) of source and load impedances are known and if they are constant, it is not difficult to design and test the most cost-effective filter. For extreme cases, this is particularly easy. It is, then, convenient to work with the A, or cascade, matrix (Fig. 3a). Each matrix term by itself describes uniquely the performance for one of the four extreme interface conditions, e.g., $1/a_{21}$ is the transfer impedance which is sufficient to calculate the filter performance if the source and load impedances are very high in terms of the characteristic impedance of the filter. Each of the four extreme conditions, Fig. 3b, (0 = nearly short circuit, ∞ = nearly open circuit) has corresponding best filters of different order, Fig. 3c being the most primitive configuration (n = 1). Higher order filters are derived by maintaining the same total capacitance, but splitting it as shown for n = 2 in Fig. 3d (partition or division; to be discussed in more detail).

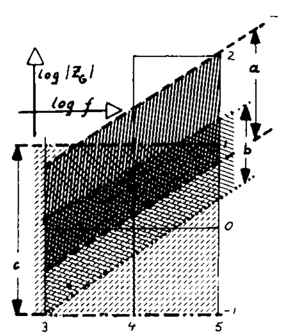


Fig. 1. Generator impedances (Z_G) of 120-V 60-Hz sources in critical frequency range of 1 to 100 kHz for normal mode *a*, common mode *b*, and highly filtered systems, common mode *c*.

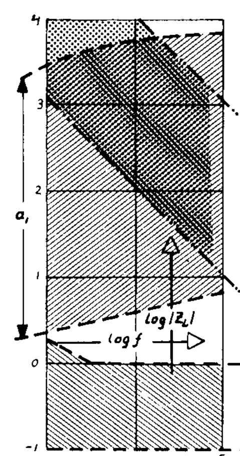


Fig. 2. Load impedances (Z_L) of 120-V 60-Hz loads, also only in critical frequency range; a_1—normal mode, except power supplies, regulated; a_2—regulated power supplies, and b—common mode.

(a)

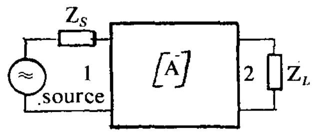
	V_2	I_2
V_1	a_{11}	a_{12}
I_1	a_{21}	a_{22}

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \bullet \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

cascade matrix of filter

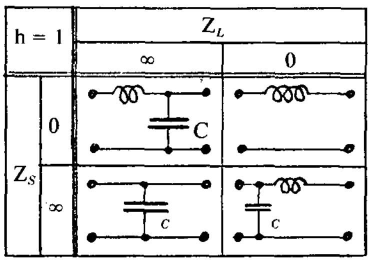
(b)

Z_S/Z_L		$Z_L = \infty$	$Z_L = 0$
Z_S	0	$0/\infty$	$0/0$
	∞	∞/∞	$\infty/0$



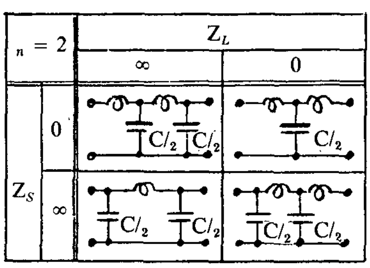
corresponding extreme $Z/_S/Z_L$ ratios

(c)



n = 1: most primitive filter configuration best suited for corresponding extreme Z_S/Z_L ratios

(d)



n = 2: next higher order of filter configuration having the same total C but much less total L. (Optimum n : 3 to 4)

Fig. 3: The two simplest classes of "best" filters for the four possible extreme cases of interface conditions

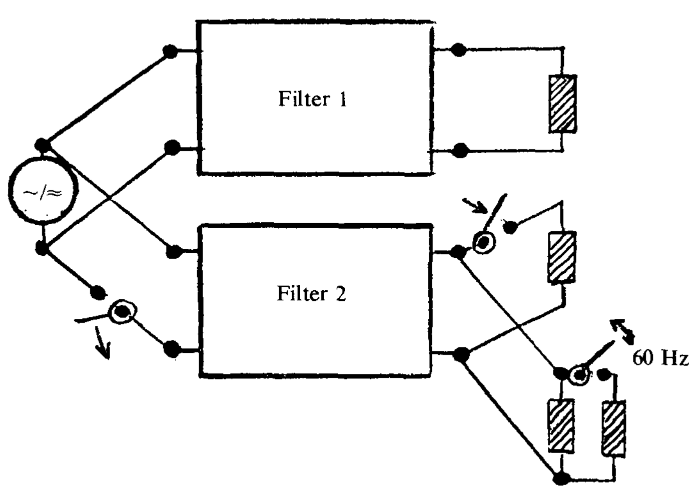


Fig. 4: The indeterminacy of interface conditions

Determinate and constant interface impedances are, however, seldom encountered in larger systems. In general, indeterminate and inconstant interfaces prevail, as the simple case of Fig. 4 illustrates. Periodic and random switching is involved on either side of the filters. (Typical examples are ships installations.) For extreme indeterminate interface conditions, which may be 0/0, ∞/∞, 0/∞, or ∞/0, all 4 filter configuations should be used simultaneously. This is not as impossible as it looks at first sight. Rather, it can quite effectively be approximated by partition, as illustrated with n = 3, for the two opposite conditions 0/∞ and ∞/0. Partitioning has two significant advantages:

(1) Very much reduced dependence on interface conditions (n = 3, 4 is best).

(2) Very much reduced inductance (reduction to about 6% for n = 2 and to 2% for n = 4, if 60 dB is stipulated for 150 kHz).

That means a drastic reduction in inductor size. In particular, in view of the ever-present 60 Hz current bias, the great reduction in size and cost also far outweighs the additional cost for more parts and assembly.

Hence, it is rather easy and expedient to insure good stop band performance with partitioning, even if the interface impedances are not known.

The negative insertion loss (already obvious in Fig. 5) and the danger of ringing, require a careful theoretical investigation. Here, the data base is of great significance in that one does not have to take into account any imaginable or high-Q interface impedance, but can confine the considerations to the clearly-defined ranges of real-world impedances. On the other hand, it is not proper to design and test filters for 50-ohm impedances only. Inspection of Fig. 5 may help to understand the severity of deficiencies inherent in MIL-STD-220A. This simple example is based on real interfaces and shows a clustering that can be explained by the extreme value theory (to be derived shortly). Such a theoretical investigation of realistic extreme value behavior is not made for the sake of theory. It is to understand a situation that could have seemed hopelessly complex and unpredictable.

Before going into the theory-made-practical stage of this article, some brief comments on ringing and insertion gain-about which widespread confusion exist, should be made. Both are oscillations caused by the high Q of the filter. In the case of ringing, the energy is applied in a step function (switching) and the oscillation has decayed before the next step function (excitation) is applied (Fig. 7). The theory of Fourier Transforms indicate that a doubling of the switched amplitude can occur. Hence, if a 120 V, 60 Hz power line is switched at the crest, the peak response may be close to 340 V. In the case of insertion gain (negative insertion loss), the energy of a harmonic is supplied continuously causing the filter output, at resonance frequency, to increase until it is Q times the input value. Q's of filters may be so high that an insertion gain of 35 dB may be realized, e.g., may magnify a 3 volt harmonic to practically 100 volts.

Such resonances (decaying or sustained) of power feedline filters happen between 1 and 100 kHz. They are of two kinds:

Eigen resonances: Conditioned on open and/or short circuits of the filter. Hence, their Q is determined by the filter alone.

Interfacial resonances: Filter and interfacial impedances form resonant circuits.

Since the interface impedances have low Q, such interfacial resonances can be disregarded in the first approximation. For filters having large capacitances, interfacial resonances resulting in low insertion gain (<10 dB) can occasionally occur at low frequencies. There are two reasons to prefer small capacitances. There is less 60 Hz reactive current and interfacial resonances are shifted higher where they can be dampened more easily and together with the eigen resonances. Practically negligible insertion gain can be expected by partitioned filters having small total capacitance in the order of $1\mu F$, as long as the eigen resonances are eliminated by damping.

Consequently, in the pass band, the key problem is eigen resonances. If the four sets of extreme interface conditions (Fig. 3b) are entered in the insertion-loss equation

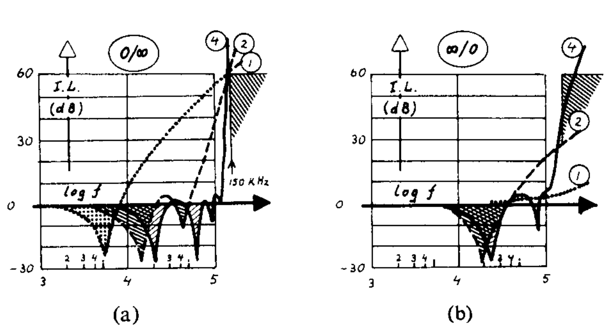


Fig. 5: The effect of partition: the most primitive filter is most sensitive to changes of interface impedances; the higher n, the less sensitive. (For stopband behavior only!)

a surprisingly transparent result, tabulated in Fig. 9, is obtained. As already seen, each of the four matrix terms a_{mn} is dominant for a corresponding particular interface condition. But, now there is a significant distinction to be made. For the *odd* cases (0/∞ and ∞/0), the insertion loss is determined by the log of a_{11} or a_{22}, respectively (short-circuit current ratio or open-circuit voltage ratio). Since the a_{11} and a_{22} (for the lossless case) have zeros (eigen resonances) and the log of zero is $-\infty$, the filters have negative insertion loss, or insertion gain, at these zeros. In contrast, for the *even* extremes, (0/0 and ∞/∞), we have to take the log of $a_{12/0}$ and $a_{21}.\infty$ respectively, which are not zero.

Fig. 9b is a sketch of what was just found, illustrated for n = 3. (Refer also to Fig. 6.) These important relations (odd ratio: insertion gain; even ratio: no insertion gain, dips in the insertion loss curves) are independent of the filter configuration and are only caused by a high Q of the filter. Hence, the filter must be lossy in the upper pass band where the eigen resonances are. In contrast to the stipulation made by proponents of lossy filters, there is no need for introducing losses in the stop band where no resonances can be imparted by interaction of filter elements. (This statement does not apply to a totally different kind of self resonances, namely the resonances of individual filter elements on account of distributed parameters. Such intra-component resonances in properly designed filters are so high in the stop band that they can be eliminated without necessarily introducing losses; whereas, the intercomponent resonances causing pass-band insertion gain must be dampened by losses.)

Losses can, for instance, be introduced in the form of magnetic or eddy current losses, by resistors in shunt with all or part of the inductors, or by resistors in series with divided capacitors. As partition shifts the eigen resonances higher in the pass band, the damping can be made so selective that the efficiency at the power frequency is not affected.

A summary of the above investigation of LC-filters operating under indeterminate interfaces can be stated on the following guide lines: (a) Partition by n = 3 or 4; (b) use low capacitance values; (c) dampen in neighborhood of eigen resonances; (d) design the filter that detrimental current bias effects are avoided.

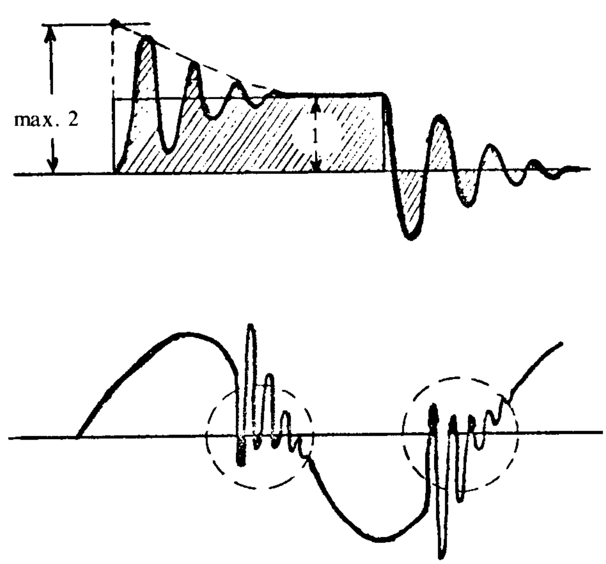


Fig. 7: Upper: ringing of D.C. pulse for Q>>1
Lower: ringing superimposed on A.C., (LC filter for SCR bridge)

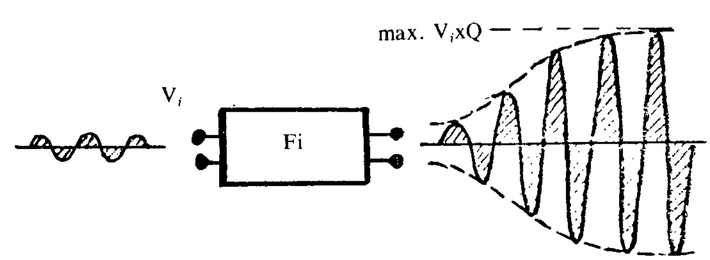


Fig. 8: Insertion gain (negative I.L.) caused by exciting resonance of high-Q filter.

(a)

I.L. = 20 Log x		Z_L: ∞	Z_L: 0
Z_S	0	$x = a_{11}$	$a_{12}/0$
	∞	$a_{21} \cdot \infty$	$x = a_{22}$

(b)

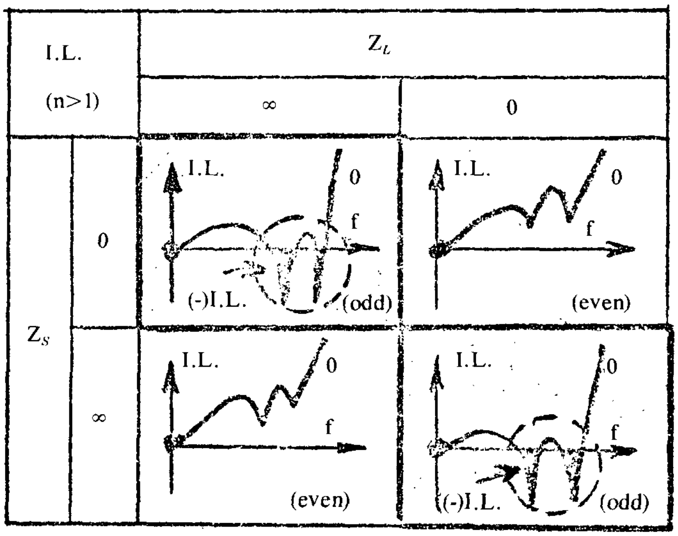


Fig. 9 The critical pass band behavior: only for odd interface conditions do eigen resonances result in insertion gain

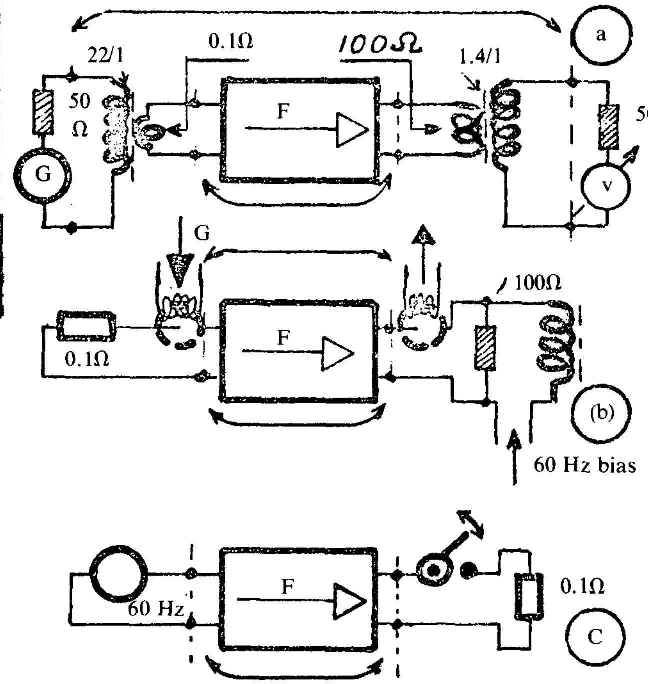


Fig. 10 Several versions of worst case measurements.
(a) 0.1/100 Ohm (& reverse !), using wide-band impedance trafos
(b) 0.1/100 Ohm (& reverse !), using ohmic resistors and current probes
(c) a simple test for ringing

Testing

Realistic worst-case testing is suggested to be as follows: Measure the insertion loss of the filter from 1 kHz to at least 300 kHz in a 0.1/100 ohm (Z_G/Z_L) system and its reverse, 100/0.1 ohm. For frequences above 300 kHz being sufficiently high in the stop band, the 50/50 ohm method, or equivalent, will do to weed out filters having undesirable resonances of individual filter elements.

Fig. 10 shows several specific realizations of the 0.1/100 ohm - 100/0.1 ohm method. In Figure 8a, broadband transformers to utilize impedance transformation of standard 50-ohm equipment. For Figure 8b, use real 0.1 and 100 ohm resistors in conjunction with current probes. With a single choke, the filter can simultaneously be tested for operation under bias conditions. Figure 8c shows a simplifed ringing test.

Outlook

The author was asked by the U.S. Air Force, Navy (in conjunction with NATO standards), and by the U.S. Army (responsible for updating MIL-STD-220A) to comment on present filter standards and to make practical, constructive proposals for correction. References 1 and 2 are the results of the author's cooperation with these agencies. Reference 2 contains a comparison with a "competing" British proposal which is theoretically intriguing; but, seemingly restricted to aircraft systems. CISPR (see separate article) has both the British and the U.S. proposals under consideration in tentative standards submitted to its world-wide committee for member comments. Although the ground work has been completed for an overdue correction of outdated filter design, it may take some time before the various agencies complete their official documents.

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2. "Indeterminate Interfaces", H. M. Schlicke, 19th Midwest Symp. on Circuits and Systems, Aug. 76 (summary only; full paper available from author, send stamped envelope.)
3. "Applied Principles of Cost-Effective Control of Electromagnetic Interference and Hazards", H. M. Schlicke; book in preparation, prospective publisher: Artech House.

This short paper was extracted by Dr. H. M. Schlicke, Interference Control Co., from his forthcoming book APPLIED PRINCIPLES OF COST-EFFECTIVE CONTROL OF ELECTROMAGNETIC INTERFERENCE AND HAZARDS which will be the main text of projected EMC courses sponsored by the IEEE.

THE ABSORPTIVE FILTER

All magnetic and dielectric materials have electrical losses when intercepting electrical and magnetic fields. These losses, which an engineer normally tries to minimize, are a function of the atomic and molecular structure of the materials. If these phenomena could be obtained at will, with a predetermination of the frequency spectrum where they occur, a valuable means of absorbing selective portions of the frequency spectrum would be available then.

Instead of using reactive elements for conducted interference reduction purposes, a direct approach of attenuating the noise currents flowing through a conductor can be realized. This attenuation is based on the phenomena of magnetic and dielectric losses. Conducted interference suppression is thus based on absorption instead of attenuation by reflection. Semiferrous and ferromagnetic materials are composed of elementary domains, spontaneously magnetized to saturation as little semipermanent magnets. When not subjected to an external field, these materials are oriented aimlessly. Between adjacent domains, there is a definite variation of the magnetization vector which depends on the anistrophy of the domains, provided the form is small. The domains are separated by limits called *Bloch walls,* which are the surfaces of the particles oriented at 90° and 180°. The magnetic flux travels essentially in closed loops within the crystal, and maintains it in a state of minimum potential energy.

The crystal itself does not produce any significant external magnetic field, except for some small fields at the ends of the Bloch walls. In these walls, the magnetic vectors gradually change direction, going from one domain to another. The finite thickness of the wall is determined by the balance of the two opposite forces: 1) the exchange energy composed in the most part by the interaction of parallel spins or magnetic vectors oriented in opposite directions; and 2) the energy of anistrophy due to the fact that a magnetic crystal produces different magnetization forces for different magnetization directions. In other words, there is an easy magnetization direction and a difficult magnetization direction. When the crystals are placed into a magnetic field, the Bloch walls move in such a way as to realize a volume increase in the easily magnetized direction and in the direction of the applied magnetic field. The magnetic vectors will then have a tendency to orient in the direction of the applied field. To describe this effect in a more theoretical manner, we can assume, initially, the variation of the initial permeability (unsaturated) at a low frequency:

$$\mu = \mu' - j\mu''.$$

The μ is a real term describing the magnetization equivalent to the permeability of the crystal. The μ' and the complex term j describe the loss due to the phenomena of wall displacement and magnetization rotation in the crystal. With a low-frequency applied field, μ'' is negligible, thus

$$\mu = 1 + 4\pi M_s^2\left(\frac{1}{\alpha d} + \frac{2}{3M_s He}\right)$$

where α represents the rigidity of the Bloch walls and d the dimension of the domain which describes the crystalline anistrophy and extends between H_{anis} and $H_{anis} + 4\pi M_s$. The shape factor H_{anis} is the equivalent anistrophy field and is defined by $4K_1/3M_s$, where K_1 is the first-order constant of anistrophy.

Therefore, the first term represents the contribution of the walls of the field. The permeability is more important since the walls are rigid and the domains small. The second term denotes the contribution due to domain rotation or magnetic vector spins. Since the applied field increases in frequency as the result of motion friction, additional losses due to resonances and reflections provide a maximum in the μ'' spectrum. These frequencies are, in general, greater than 1 MHz. The increased loss due to resonance and the movement of the Bloch walls is given by

$$\omega_1 = \delta(8\pi\alpha)^{1/2}\left(\frac{A}{K}\right)^{1/4}$$

Where δ is the gyromagnetic ratio and A is the exchange energy per volume unit. The resonance condition, for pure domain rotation, is given by

$$\omega_2 = \delta He = \frac{8}{3}\left(\frac{\pi\delta M_s}{\mu' - 1}\right).$$

For very high frequency, both losses decrease with no external permanent magnetic field. These would be added to He and become very small at the maximum value of He:

$$\omega_{limit} = \delta He_{max} = \delta(H_{anis} + 4\pi M_s).$$

Above the limit, there are negligible magnetic losses. The dielectric material will also have resonance effects at high frequencies which introduce dielectric losses:

$$\xi = \xi' - j\xi''.$$

That is, the permittivity or the complex dielectric constant varies with frequency. Due to ionic conductivity (free carriers), ξ'' will vary considerably in the low-frequency spectrum:

$$\xi'' = \frac{\sigma}{\omega}$$

where σ is the conductivity of the dielectric material. As frequency increases, the losses due to ξ'' of the Maxwell-Wagner type teach a maximum in the dielectric material due to interfacial polarization. In fact, if the dielectric is composed of particles or grains in a medium which has a different electrical characteristic, then resonant frequency ω' decreases, ξ' decreases, and ξ reaches a maximum:

$$\omega_1 \leqslant \frac{\sigma_1 d_2 + \sigma_2 d_1}{\xi_1{}^1 d_2 + \xi_2{}^1 d_1}$$

where σ_1 and σ_2 represent the conductivity, $\xi_1{}^1$ and $\xi_2{}^1$ the permittivity of both dielectric mediums, and d_1 and d_2 the dimensions of each of the dielectrics. Ferrites also display a dielectric constant, and thus the material absorbs the electric field about the conductor imbedded in the semiferrous material.

The absorptive filter is shown in Fig. 1 and the Capcon ad. A conductor is coiled into a helix and imbedded into the absorptive material. A dielectric sleeve separates the absorptive material from the case and provides additional distributed capacitance to further increase electric field losses. Due to the reduction in the flux field lines between the helix turns by the absorptive material, the Q of the helix is reduced to unity or less than unity. Since the Q is either unity or less than unity, the helix does not represent an inductance, but will appear essentially as a series resistance which is proportional to the square of the frequency of the current flowing through the conductor.