THE IMAGE PARAMETER NETWORK

The low-pass network is the most widely used type in reducing conducted electromagnetic interference, or noise. The most generally available filter type is the pi-type network. The pi network consists of a series indicator and two shunt capacitors in a three-branch filter. Filters of this type, designed for shielded room use, have as many as seven branches. Figure 1 illustrates the construction of a typical three-branch pi network.

The two feedthrough capacitors are mounted on cups which make peripheral contact with the filter case. It is assumed that a cylindrical case is used. The cylindrical configuration, as compared to square, rectangular, or other physical configurations, provides the highest order of attenuation with minimum perturbation throughout the guardband (stopband) spectrum of the filter.

To reduce I^2R losses, the series inductor is usually wound on either a ferrite or permalloy core to increase inductance. For purposes of reducing the indicator's external field, the core is usually toroidal.

The capacitor cups, in addition to peripherally terminating the feedthrough capacitors, provide internal shielding, resulting in additional internal isolation between the input and output terminals of the filter assembly. Since such care is taken in the assembly of the filter to provide input-output isolation, it follows that the same attention should be given to providing external isolation. Realizing the attenuation a filter is capable of providing, input-output isolation becomes very critical. Either a bulkhead that is sealed against radio frequency energy penetration or a shield plane of at least one-sixteenth wavelength ($\lambda/16$) in both the length and width dimensions of the filter's cutoff frequency is required to provide the necessary external isolation.

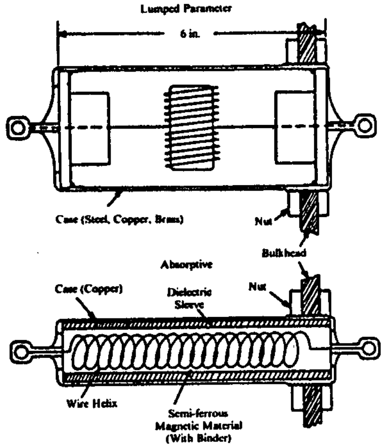

Figure 1. Typical Interference Reduction Filter Construction.

In general, the image parameter filter network can be designed to Butterworth or Chebyshev functions, although most standard filters are designed to the Chebyshev function. The expression $A(\omega) = 10 \log [1 + f(j\omega) f(-j\omega)]$ describes the attenuation in dB for both the Chebyshev and Butterworth functions in the frequency variable ω. These functions will contain powers of ω from 1 to η, the highest power being equal to the number of branches which will be contained in the resulting ladder network. The ladder network is realized algebraically from the above transfer function in a logical sequence of operations. First, the driving point or input impedance function is developed. Second, the ladder network is obtained by application of the pole removal process, the Cauer synthesis, or another applicable method.

Chebyshev filters have a passband response that limits the ripple to a predetermined envelope. This is accomplished by arbitrarily assigning a constant multiplier to the Chebyshev polynomial. Ripples from less than 1 to more than 3 dB are common. In general, the greater the passband ripple, the more rapid the stopband roll-off, which can be described by a monotonically increasing function. The number of ripples in the passband are directly related to the number of branches in the ladder network, or the "degree" of the Chebyshev polynomial.

Butterworth filters, on the other hand, have the characteristic of being "maximally flat" in the passband. This means the response in the passband is the flattest possible response obtainable without changing direction. This response is monotonic, ever increasing in attenuation as frequency increases. The roll-off in the guardband is also monotonic, approaching a constant slope of 6 dB per octave per ladder network branch.

All of the above is true if the source and load impedances are equal and remain constant throughout the frequency spectrum. Where such networks are used as interference reduction filters (for example, in power line filtering), it is seldom that load and source impedances are equal and constant.

Of course, the design of filter networks is flexible with respect to termination ratios. The filter designer is not restricted only to circuits which have equal source and load impedances. Filters which have only a single-termination impedance are the easiest to obtain. A search of manufacturers' catalogs reveals that only single-termination impedance designs are available. Image parameter networks can be designed which operate either from a near zero impedance source to a finite impedance load, or from finite sources to an open circuit. In such designs, all the available energy is consumed in only one termination. Thus, networks can be designed by the circuit design engineer or the EMC/EMI engineer for any ratio or source-to-load impedance, including the most common condition of equal terminations, which is the easiest to design.

This illustrates why unequal termination networks are not designed and are not available from the major interference reduction filter manufacturers; the following changes would be required in the image parameter filter shown in Figure 1 for unequal termination impedance. If a 10-to-1 load-to-source impedance ratio is assumed, then the capacitance ratio would have to be 10 to 1 with the larger value capacitor in shunt with the lower impedance. The series inductor would have to be increased by a factor of 1.1 times inductance. This is in accordance with the Butterworth function.

An additional area that bears consideration is that reactive components change from a positive to a negative reactance, or the inverse, at some frequency, generally in the guardband area of the filter response. The larger the value of capacitance, the lower the frequency at which the capacitor reaches self-resonance and, consequently, becomes inductively reactive. Above this self-resonant frequency, parasitic resonances occur. The inductor above resonance becomes capacitively reactive and also exhibits parasitic resonances above the self-resonant frequency. Thus, in a low-pass image parameter network, the internal shields provide the primary attenuation above the frequency of self-resonance of the network's reactive components.

Phase-shift and time delay characteristics of signal-line filters are not normally taken into consideration unless the engineer has to comply with the requirements of the classified EMI specifications. Such characteristics should be taken into consideration and may be of great importance, even though the pertinent applicable specification or specifications do not delineate any requirements for phase-shift limits and maximum time delays.

Low-pass image parameter filters display a phase characteristic which is 0 degree at zero frequency and increases steadily throughout the passband. If this increase is a linear or straight-line function of frequency, the time delay through the network remains constant within the range of linearity. In the guardband, phase characteristics become very erratic and time delay varies widely. Phase shift usually approaches a constant value (which is a multiple of 90 degrees, depending on the number of branches) and the delay also approaches zero.

In data transmission systems, where filters may be required to pass modulated carrier tones, the phase-shift and envelope-delay behavior is of considerable importance. The degree of phase-shift linearity will affect the amount of signal distortion. Filter design for a constant time delay can be obtained through the use of Bessel polynomials. The resultant networks display exceptional linear phase characteristics well into the stopband, but they have very poor response behavior. In some cases, a phase correction network added to the standard network will provide the required response characteristics while limiting phase-shift nonlinearities. Obviously, this compensating network adds materially to the size, weight, and cost of the filter assembly.

When an image parameter filter is required to pass a step function, such as low-frequency square waves or transients, the response and phase characteristics of the filter combine to alter the form of the signal by introducing overshoot and ringing at the load. Overshoot may exceed 25 percent for some types of low-pass filters. Some improvement can be realized by phase compensation, but the engineer will have to accept some degree of overshoot and ringing.

THE ABSORPTIVE FILTER

It has been known for some time that 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 these 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 then be available.

Instead of using reactive elements for conducted interference reduction purposes, a direct approach of attenuating the interference and 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.

The absorptive filter is shown in Figure 1. 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.

A TYPICAL FILTER APPLICATION

Interference reduction filters are most often used to reduce conducted spurious energy in power circuits, both ac and dc. The current requirements for an operating circuit are seldom static but are dynamic in that current requirements will change very rapidly.

In combination with an initial current demand by the circuit in use, the external flux field around the conductors carrying power and current is higher than that of signal leads. Thus, the spurious energy creating the flux field has to be reduced to minimize coupling into sensitive circuits. Spacing between power carrying conductors and signal conductors will reduce the coupling factor due to transmission losses through space. Advantage can be taken of the high space transmission losses in the vorticity field area. Twisting of the leads will also reduce the external flux field about conductors, thus further inhibiting the coupling of any spurious energy into adjacent susceptible circuitry. However, even though advantage can be taken of these techniques, the use of interference reduction filters provides the greatest reduction of spurious energy transmission over wires, providing proper filtering techniques and procedures are used. If not, additional interference voltages will be generated by the filter networks, while only minimum attenuation of the spurious energies are obtained.

Due to the rapid transfer time from the conducting to nonconducting state of solid-state rectifiers, power supplies couple rapid current changes back into ac power lines. Such rectifiers are conducting over only a very small portion of the sine wave current source. Due to the dc back bias of the developed dc power supply potential, the rapidly changing reflected load is seen through the power transformer by the interference reduction filter network. The changing impedance will alter the operation of the filter network, particularly the image parameter network, creating additional spurious energies and, in some cases, cause electronic system performance degradation.

To examine this effect in more detail, refer to Figure 2 which shows an ac source consisting of a stepdown transformer of the power distribution type furnishing 300 kVA. Four types of filters will be examined for their operating characteristics in this application, which is a typical application condition. The first filter type is a pi type of image parameter low-pass network using lumped constants, the second filter network is a T type of image parameter low-pass network, again using lumped constants. The attenuation of both these networks, if tested in accordance to MIL-STD-220A, would be the same. The third interference reduction filter network consists of an absorptive filter and a $1 \mu F$ capacitor. The fourth filter is the same absorptive filter, but without the lumped capacitor. The attenuation of these networks, when tested to MIL-STD-220A, is comparable to the image parameter networks.

The load circuit is represented by the transformer, either stepup or stepdown, and its secondary load. The secondary load varies from 0 to 1×10^6 ohms, depending on the conducting or nonconducting state of the rectifiers that are connected to the secondary winding of the power transformer. The nominal current input into the transformer is 10 amperes. Thus, the nominal load impedance would be 12 ohms (if considered to be resistive), and the source impedance is approximately $100 M\Omega$, neglecting the series impedance of any power distribution wiring between the power distribution transformer and the filter network.

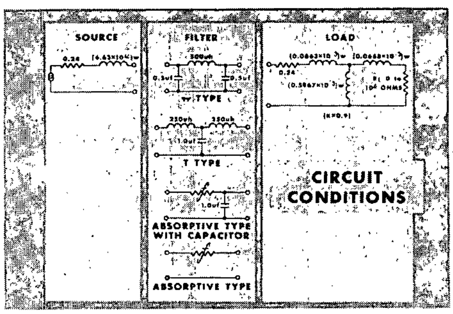

Figure 2. Circuit Conditions.

Since the absorptive type of interference reduction filter appears as an equivalent series resistor, a test method was developed to determine this series resistance directly. The test is the admittance-transfer method, and Figure 3 illustrates the series resistance of absorptive-type filters obtained from two vendor sources. By using this test method, similar data can also be obtained for the image parameter network.

The resonant frequency of an image parameter filter will occur at the filter's cutoff frequency, that is, the frequency at which the network will begin to exhibit attenuation to that frequency. Thus, the pi network, shown in Figure 2, with component values of $0.5 \mu F$ shunt capacitors and a $500 \mu H$ series inductor, will resonate at approximately 8000 Hz, as shown in Figure 4. As shown in this figure, the gain of the filter will increase as the reflected load of the rectifiers through the power transformer decreases, or in other words, as the secondary of the power transformer in the power supply sees an open circuit. The frequency at which the gain occurs also shifts to a lower frequency due to reduced current flow through

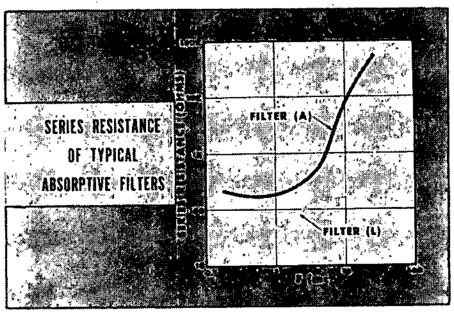

Figure 3. Equivalent Source Dissipative Resistance.

the series inductor of the pi-type filter network, resulting in a higher series inductance.

The series inductor is wound on a permalloy core with a nominal permeability of 125. When the secondary load resistance is equal to $1 M\Omega$, the gain reaches 27.6 dB at about 7000 Hz. For filters with other values of lumped constants, the resonant frequency and the gain will be different. This figure also shows that 12 ohms is approximately the matching load impedance for this pi-type image parameter network. The resultant oscillation at the resonant frequency will, of course, appear in the power line and in the power supply, and can cause additional interference problems when using electronic or adjacent electronic equipment. The oscillation will be 20 Hz of a damped oscillation at the resonant frequency, damping out after $333 \mu S$. There will be a 98 percent initial overshoot at $1 M\Omega$ reflected load.

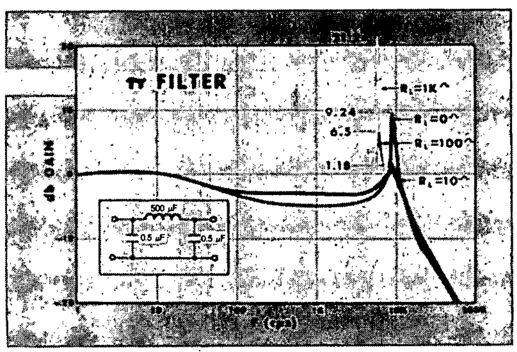

Figure 4. Filter Characteristics.

Figure 5 shows the improved operation of the T type network. The maximum gain at $1 M\Omega$ reflected load will be 7.1 dB as compared to the 27.6 dB of the pi network. Both networks, that is, the T and the pi, will exhibit the same attenuation when tested in accordance with MIL-STD-220A with 50-ohm resistive source and load resistances.

As in the previous figure, the optimum load is approximately 12 ohms for the T network. However, it must be borne in mind that the power transformer and the power supply will vary widely in impedance over the

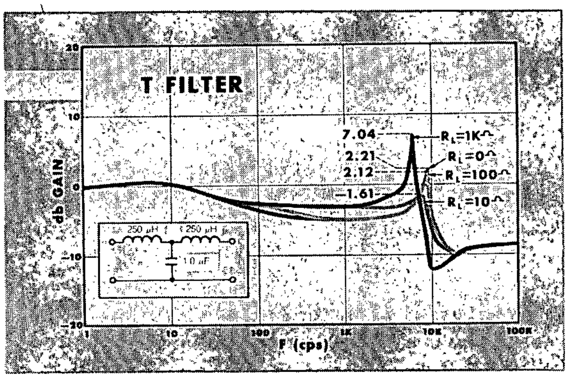

Figure 5. T Filter Characteristics.

filter's passband and guardband portions of the frequency spectrum. Thus, the 12 ohms indicated is the resistive equivalent at the power line frequency of 60 Hz. The change in source and load impedance at frequencies other than the power line frequency is not considered in these figures, but should be taken into consideration in the performance characteristics of a filter network.

Figure 6 shows the gain and resonant frequency of the absorptive-type filter with a 1-μF shunt capacitor. The resonant frequency does not change as in previous figures of the image parameter networks. The gain shown is attributable to the resonance circuit comprised of the 1μF capacitor and the power supply power transformer. When the reflected load is less than 50 ohms, the network exhibits attenuation.

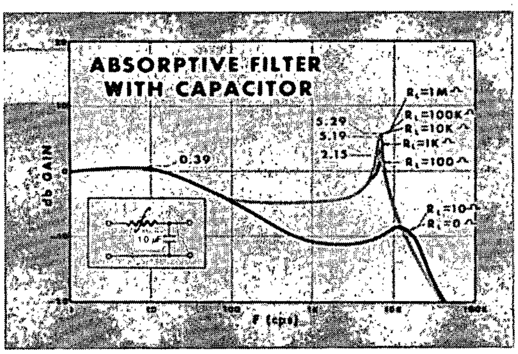

Figure 6. Gain and Resonant Frequency of the Absorptive-Type Filter with Shunt Capacitor.

When the shunt capacitor is removed, the transformer capacitor resonant condition no longer exists, as shown in Figure 7.

Figure 8 illustrates the test configuration for determining the admittance-transfer of a filter network, be it image parameter or absorptive. Rated current, either ac or dc, can be introduced across the filter to determine

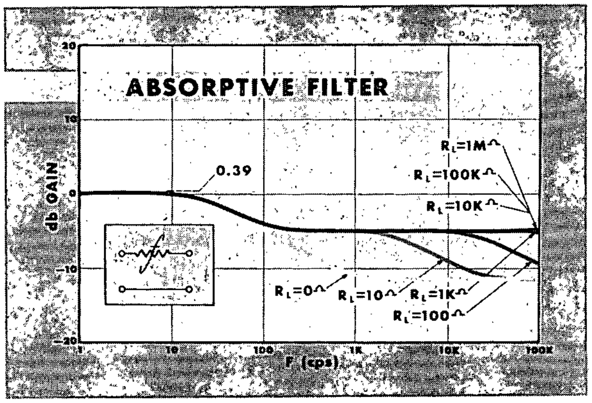

Figure 7. Gain and Resonant Frequency of the Absorptive-Type Filter without Shunt Capacitor.

the change in operating characteristics at rated currents. The load resistor prevents the short-circuiting of the shunt capacitor of a π network and, thus, should be equal to the expected impedance of the load at the power frequency or nominal impedance of the signal line. The error introduced by this resistor is small. The series resistance of the filter obtained from this test can be used to compute the attenuation at the frequencies of interest in the using circuit.

The series resistance of the filter is obtained by dividing the voltage obtained from the current probe into the voltage across the filter under test as obtained from the signal generator. In use, the frequency variable series resistance of the filter network and the load impedance will act as a voltage divider, and thus, if in the equation the load resistance is made to equal 30 ohms, the data obtained from MIL-STD-220A can be correlated:

$$\alpha = 20 \log \frac{R_i + R_L}{R_L} \qquad (9)$$

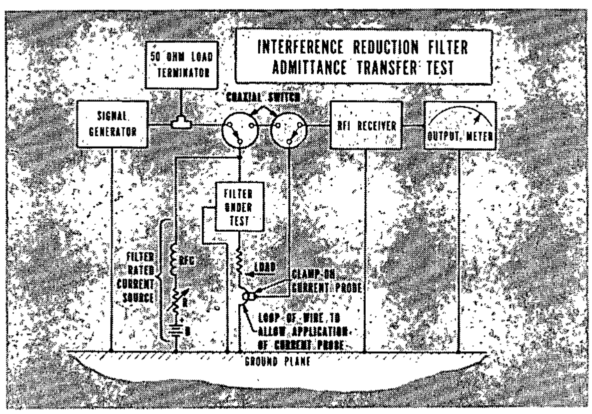

Figure 8. Interference Reduction Filter Admittance Transfer Test.

This article was written for ITEM 85 *by Dr. Anthony W. Laine, Rockford, IL.*