Microwave Measurement of Stripline Filter Characteristics Using Time Domain Techniques

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ABSTRACT.

This paper describes the technique for obtaining the frequency response (0.1 - 8.0 GHz) of a microwave stripline filter using an automated network analyzer (ANA). Unwanted reflections caused by the coaxial to stripline discontinuities are mathematically removed using time domain gating. A versatile 50 Ω microwave stripline fixture was designed and built; then, a low pass filter was designed and applied to the stripline; finally, the filter response was measured using the techniques presented in this paper. For comparison, the designed filter response was calculated analytically using a transmission line model.

I. INTRODUCTION.

The frequency and time domain stripline responses presented in this paper were obtained using a Hewlett Packard 8510 Automated Network Analyzer (HP8510 ANA). This analyzer is capable of making complex reflection (S11) and transmission (S21) measurements in the frequency domain and with on board digital signal processing, is capable of executing a Fourier transform to express the unit impulse response in the time domain. [1]

The major advantage the HP8510 over earlier ANA's, for example the HP8409, is the on board signal processing. The HP8409 was limited to making measurements only in transmission lines where good standards were available (i.e. 50 Ω coaxial lines). For example, if stripline devices were to be measured, it would be necessary to isolate the stripline response from the device in the stripline. To accomplish this, the designer had two choices: either construct accurate calibration standards for the given stripline or make excellent (>30 dB return loss) transitions from the standard coaxial line to the stripline. Both options are expensive and time consuming.

The HP8510 can solve the non-standard transmission line problem with its time domain signal processing. This is accomplished by transforming the frequency domain data to the time domain. Then, knowing the approximate electrical spacing between the unwanted transition responses and the embedded device response, a gate can be placed in time where the device response is originating and the responses outside the gate can be omitted. This gated data can be transformed back into the frequency domain to represent the spectral response of the device relative to the non-standard transmission line.

II. STRIPLINE DESIGN.

The stripline was designed to have a characteristic impedance of 50 Ω and to operate in the frequency range of 0.1 to 8.1 GHz. The stripline was constructed in five layers. The outer conductive layers were made from 1/16-in aluminum sheet metal. The inner insulating layers were made from 1/8-in G-10 (resin filled woven glass), which has a relative permittivity of 4.2. The center layer is a sheet of single sided 1-oz copper clad 20 mil circuit board. The circuit board was etched leaving only a 117 mil center conductor, which was the 50 Ω design width calculated using design curves. [2] The length of the stripline was 24 in, which was long enough to provide adequate time domain gating; the total width of the stripline was 11 in which was sufficient to restrict the field to the interior and thereby avoid fringing.

Two commercially available N-female bulk connectors were used to interface the stripline to the coax. These connectors had brass center conductors and 3/8-in Teflon jackets. The Teflon jackets were machined flat on both the top and the bottom to accommodate a 0.270 in interior stripline spacing. Slots were cut in the brass center conductors to allow the center strip on the printed circuit board to slide into it. This connection was then carefully soldered.

The top plate and the top G-10 layers were not permanently bonded, but were held in place with clamps for easy exposure to the center conductor. Filters could then be applied to the stripline with copper tape patterns attached to the center conductor.

III. CALIBRATION AND MEASUREMENT OF BASELINE STRIPLINE.

The designed stripline was a symmetrical two port device. The choice of the input and the output ports was arbitrary. A input port was selected and designated as "port #1", and the remaining
port, the output port, identified as "port #2". The HP8510 was programmed to perform a full one path calibration from 0.1 to 8.1 GHz in 400 frequency steps. This produces a step frequency, $F_{\text{step}}$, of 20 MHz. Consideration of the step frequency is important since it sets the alias free time interval for the unit step response. Using equation (3.1), the alias free time interval, $T_s$, is found to be 25 ns.

$$T_s = \frac{0.5}{F_{\text{step}}} \quad (3.1)$$

The alias free distance, $D_s$, was the distance where the time domain response repeats and was determined using equation (3.2).

$$D_s = \frac{c}{T_s} \quad (3.2)$$

where $c = 2.998 \times 10^8 \text{ m/s}$.

$D_s$, calculated at 3.66 meters for the designed stripline, was more than adequate in measuring an echo free response of the 0.61 meter long stripline.  
The calibrated transmission path started at the HP8510 port A and traveled through a three foot APC-7 to N-male 3.5 mm semi-rigid cable, into a N-female to N-female "bullet" adaptor, through another three foot N-male to APC-7 3.5 mm semi-rigid cable, and terminated at the HP8510 port B.  
The calibrated reflection phase reference was done with N-type standards (short, open, and broadband load) at the end of the first three foot cable.  
The N-female "bullet" was removed from the transmission path and the stripline was inserted in its place. $S_{11}$ and $S_{21}$ were measured, and the unit impulse response obtained for the baseline stripline; figures 3-1 and 3-2 show the responses for $S_{11}$ and $S_{21}$ respectively.

The y axis for all plots presented in this paper represents a magnitude response, in decibels (dB), relative to a short circuit (0 dB) with the y axis scale equal to 10 dB/division. The x axis, representing time, begins at -1 ns and stops at 24 ns with each horizontal division representing 2.5 ns. The 25 ns time base span allows for the entire alias-free region to be shown.

The time domain plots provide insight into where the unwanted stripline effects are dominant. Using equation (3.2) it can be shown that each 2.5 ns division accounts for a distance of 0.366 m (1.201 ft). Since $S_{11}$ represents reflection, this distance accounts for two way travel along the stripline. This implies that the location of the reflections occur at half the distance calculated from equation (3.2). Therefore, for reflection, approximately 2.5 ns is equal to a 0.6 ft distance traveled in the stripline.

Figure 3-1 shows the first dominant reflection at 0 ns and a secondary reflection approximately 7.5 ns later. These reflections correspond to the locations of the coaxial to stripline transitions. In between the transitions is the "quiet" interior region of the stripline, and this region will define the time domain gate. The response outside the gate is mathematically forced to zero (−∞ dB) when the gate is used, and the gated frequency response is obtained by transforming the gated time domain back into the frequency domain. Figure 3-3 shows the stripline frequency response with and without the gate. The x axis scale for all frequency plots in this paper starts at 0.1 GHz and stops at 8.1 GHz with each horizontal division representing 800 MHz. Figure 3-3 shows an improvement in the stripline mismatch because of gating of approximately −10 dB in the lower part of the frequency band and about −25 dB in the upper portion of the band.
Transmission time domain data must be interpreted differently. Figure 3-2 shows the transmission time domain. The first dominant transmission signal is delayed 3.9 ns. This delay represents the first dominant signal delayed, via the electrical length of the stripline, with respect to the calibrated transmission path. Figure 3-2 shows some additional signals in the transmission time domain which are both reduced in magnitude and delayed longer in time than the principal path. These other signals represent the waves that traveled through the stripline more than once; hence they are referred to as "multi-path signals." These multi-path signals occur because of the mismatch of the coaxial transitions at both ends of the stripline causing a wave to reflect back and forth along the stripline. These signals are obviously delayed in time from the principal path. The transmission gate is therefore centered around the principal path as shown in figure 3-2.

The consequence of the multi-path signals, shown in the ungated frequency response of figure 3-4, is the spectral ripple which is most pronounced at the high frequencies. The gated frequency response, also shown in figure 3-4, illustrates the elimination of the unwanted spectral ripples.

**IV. LOW PASS FILTER DESIGN.**

A multi-element low pass stripline filter, shown in figure 4-1, was designed to have a corner of approximately 1.0 GHz. The wider 1.0-in sections represent capacitive elements with respect to the nominal 50 Ω stripline width of 0.117 in. This geometry was easily fabricated by applying copper tape, with conductive adhesive, cut to designed lengths, over the permanent 50 Ω stripline center conductor.

The 0.117-in center conductor sections were designed to have a characteristic impedance of 50 Ω as previously mentioned, while the 1.0-in center conductor sections had a characteristic impedance of 11.5 Ω, as obtained from the reference material. [2] A program in C was written which calculates $S_{11}$ and $S_{21}$ for a stripline filter of known characteristic impedance sections. This was accomplished using the transmission line equation: [3],

$$Z_{in} = Z_{r} \frac{\cos(2\beta x) + j Z_{o}/Z_{r} \sin(2\beta x)}{\cos(2\beta x) + j Z_{r}/Z_{o} \sin(2\beta x)}$$

(4.1)

where $Z_{in}$ is the input impedance of a transmission line of length $x$, of characteristic impedance $Z_{o}$, and terminated with a fixed impedance, $Z_{r}$. The propagation factor, $\beta$, for the TEM mode is related to the operating frequency, $F$, by,

$$\beta = \frac{2\pi F}{c}$$

(4.2)

The program initializes $Z_{r}$ to 50 Ω, $Z_{o}$ to 11.5 Ω, and $x$ to 0.25 in. Using equation (4.1), the input impedance of the first stage is calculated. This impedance becomes the fixed load impedance of the second stage. The second stage input impedance is again calculated using equation (4.1) with $Z_{o}$ set to 50 Ω and $x$ to 1.0 in. This cascading procedure is repeated through all five stages of the filter until the final value of $Z_{in}$ represents the input impedance of the entire filter, $Z_{r}$.

The complex reflection coefficient, $\Gamma$, with respect to the 50 Ω stripline is then calculated using the equation

$$\Gamma = \frac{Z_{r} - 50}{Z_{r} + 50}$$

(4.3)

$S_{11}$ and $S_{21}$ are next obtained from equations (4.4) and (4.5), respectively.

$$S_{11} = 20 \log |\Gamma|$$

(4.4)

$$S_{21} = 10 \log (1-|\Gamma|^2)$$

(4.5)

Using this analysis, the predicted response for the filter shown in figure 4-1 was obtained; see table 4-1.

Figures 4-2 and 4-3 show the measured unit impulse responses for the filter inside the stripline. The $S_{11}$ time domain data show considerably more response inside the previous "quiet" gated region of the stripline without the filter; see figure 3-1. This additional response arises from the reflection of the capacitive filter elements. The time domain gate used to isolate the filter reflection response from the stripline response is illustrated in Figure 4-2. This gate, which was the same as used for the baseline stripline $S_{11}$ measurement, retains the data inside the filter time.
interval and forces all outside responses to zero.

Table 4-1: Low Pass Filter Calculated Response.

<table>
<thead>
<tr>
<th>Freq (GHz)</th>
<th>S11 (dB)</th>
<th>S21 (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>-13.47</td>
<td>-0.2</td>
</tr>
<tr>
<td>0.3</td>
<td>-7.21</td>
<td>-0.92</td>
</tr>
<tr>
<td>0.5</td>
<td>-8.86</td>
<td>-0.6</td>
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<tr>
<td>0.7</td>
<td>-25.8</td>
<td>-0.01</td>
</tr>
<tr>
<td>0.9</td>
<td>-24.14</td>
<td>-0.02</td>
</tr>
<tr>
<td>1.1</td>
<td>-5.49</td>
<td>-1.44</td>
</tr>
<tr>
<td>1.3</td>
<td>-0.65</td>
<td>-8.59</td>
</tr>
<tr>
<td>1.5</td>
<td>-0.14</td>
<td>-14.85</td>
</tr>
<tr>
<td>1.7</td>
<td>-0.06</td>
<td>-18.89</td>
</tr>
<tr>
<td>1.9</td>
<td>-0.03</td>
<td>-21.14</td>
</tr>
<tr>
<td>2.1</td>
<td>-0.03</td>
<td>-21.76</td>
</tr>
<tr>
<td>2.3</td>
<td>-0.04</td>
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</tr>
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<td>2.5</td>
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<td>2.7</td>
<td>-0.44</td>
<td>-10.16</td>
</tr>
<tr>
<td>2.9</td>
<td>-12.56</td>
<td>-0.25</td>
</tr>
<tr>
<td>3.1</td>
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<td>3.3</td>
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</tr>
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<td>3.9</td>
<td>-0.01</td>
<td>-27.05</td>
</tr>
<tr>
<td>4.1</td>
<td>-0.01</td>
<td>-28.41</td>
</tr>
</tbody>
</table>

Figure 4-2 shows the time domain gate used to recover the filter transmission spectral response. This gate is not as narrow as the gate used for the baseline stripline where only the principal path was retained. This gate is expanded to include approximately 2.5 ns of delayed multipath signals. The wider gate is required to retain the response of the interaction of the filter elements with each other. This is the essence of how a filter works when viewed in the time domain. Each element produces a reflection vector (magnitude and phase). By controlling the capacitance and separation, these vectors add to produce the desired frequency response of the filter. The gate must be wide enough to allow for this interaction but still sufficiently narrow to reject the interaction of the filter elements with the stripline transitions. The physical size of the stripline has to be large enough to allow this precise time domain signal separation.

Figure 4-5 shows the gated retransformed filter transmission spectrum. The calculated $S_{21}$ values from Table 4-1 are plotted along with the measured data for comparison.
V. CONCLUSIONS.

A stripline was designed, built, and measured. Time domain gating was used to eliminate the unwanted coaxial to stripline transitions and thereby to provide a "quiet" region where stripline filters could be measured.

A program was written to calculate the response of a stripline filter. A low pass stripline filter was then designed, and its performance was predicted using the software. The filter was applied to the interior of the stripline and response was measured.

Time domain gating was used to isolate the filter response from the stripline transitions. The location and size of the gate used for obtaining $S_{11}$ of the filter was the same as used to isolate the previously "quiet" region in the baseline measurement. It was shown that the transmission gate had to be expanded to include the delayed response of interactions among the filter elements. The gate used was carefully selected to include the delayed filter response but to exclude the delayed stripline to coax transition responses. Careful consideration of the physical dimensions of the stripline fixture and of the filter, and the alias free distance is imperative in achieving successful signal separation.

Agreement between the predicted filter response and the measured filter is good. The measured filter appears to have a 200 MHz lower corner frequency. This is apparently because of the fringing capacitance of the capacitive filter elements, which were not modeled in the calculation software. The measured stopband is approximately 10 dB lower than calculated. This can also be accounted for by the unmodeled fringing capacitance which provides more shunt reactance at the higher frequencies as well as the copper losses of the imperfect stripline conductors. The calculation predicts a resonance at 3 GHz which happens to be the frequency where the filter elements are electrically spaced a half wavelength apart. This resonance at 3 GHz is not as pronounced in the actual measured filter because of the copper losses and the fringing capacitance.

The results in this paper demonstrate that predicted frequency response of stripline filters can be extracted from the combined responses of the filter itself and stripline fixture it is in. This is successfully accomplished by using time domain gating. It is believed that this technique can be used to measure the frequency response of other stripline devices such as amplifiers, and mixers.

REFERENCES.

