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# PRACTICAL DESIGN TECHNIQUES FOR FR4 MATERIAL-BASED MICROWAVE HAIRPIN FILTERS

Nikholas G. Toledo

Communications Engineering Division
Advanced Science and Technology Institute
C.P. Garcia Ave., UP Technopark, Diliman
Quezon City, Philippines 1101

## ABSTRACT

*This paper presents practical design techniques for building hairpin filters using FR4 laminates. This type of microwave microstrip filter is among the most popular low microwave frequency filters because of its compact size and ease of manufacturing. However, building the hairpin filter using FR4 laminates is very difficult to do because of the relatively poor performance of the laminate at the microwave region. The motivation to use FR4 in the low microwave frequencies is its availability and inexpensive manufacturing process. Methods and techniques were developed to address these problems of the FR4.*

## I. INTRODUCTION

### *1.1 Hairpin Resonator Filter*

The hairpin resonator filter is one of the most popular low frequency microwave microstrip filters. It is a variant of the edge-coupled filter but does not require as much circuit space. It also does not require critical grounding unlike the interdigital and combline filters thus making the hairpin filter easy to manufacture. Unfortunately, very little literature on the design of this type of filter exist. The synthesis and yield optimization of this filter is usually done with the aid of Electronic Design Automation (EDA) software tools. For these EDA tools to produce simulation results that reflect actual data, it is imperative that the filter structure must be modeled correctly and the substrate or laminate must be characterized precisely within the desired frequency band. Most EDAs require a single value of each laminate parameter that they will use throughout the frequency band.

### *1.2 FR4 Laminate*

The most common circuit board for frequencies below the microwave frequency is the FR4 laminate. This is the fire-retardant variant of the G10 laminate. The material's properties, like $\varepsilon_r$, *tan* $\delta$, thickness and roughness, are poorly controlled. These values vary among manufacturers and sometimes within the board itself. These properties can be ignored in low frequency applications where the board simply acts as an electronic carrier and package. At microwave frequencies though, the board becomes part of the circuit. Full characterization of the FR4 and its frequency response is very important in order to built FR4 microwave devices. The varying values of the FR4 properties between manufacturer, spatial dimensions and frequency make this very difficult to do. The varying dielectric constant, $\varepsilon_r$, makes quarter-wavelength constructions unpredictable. The high loss tangent, *tan* $\delta$, introduces unacceptable attenuation values. Beyond 3 GHz, the use of the FR4 is not anymore recommended. (There are however some amateur designs [1] that employ FR4 up to 5.7 GHz but these modules are very lossy for industry standards.) Choosing values close to the data sheet values and values derived from analyzing quarter and half wavelength microstrip lines normally result in EDA simulation results that are off the actual response by several hundred megahertz.

The FR4 however is readily available. It is processed using inexpensive techniques resulting in a laminate that is low-cost compared to most stable microwave laminates [2]. Methods were thus developed to overcome the problems of FR4. This paper presents a practical approach to designing hairpin filers. These techniques do not utilize simulation and synthesis software and are solely based on empirical results.

## II. INITIAL HAIRPIN FILTER DESIGN

The hairpin filter is a variation of the edge-coupled band pass filter. The resonator elements of the hairpin filter are bent to save circuit real estate. To do this, a sliding factor is introduced in the bending area (see Fig. 1) [3], [4]. The steps in designing a typical edge-coupled filter are enumerated in [5].

However, it was observed that the structure shown in Fig. 1c has a large insertion loss and an unacceptable return loss when built on the FR4 laminate. A tapped hairpin resonator filter was found to have a better frequency response in terms of its insertion and return losses. The tapped hairpin filter is a variant of an edge-coupled filter containing a matching stub (see Fig. 2). Varying the tapping distance and the length of the matching stub varies the return loss and consequently the insertion loss.

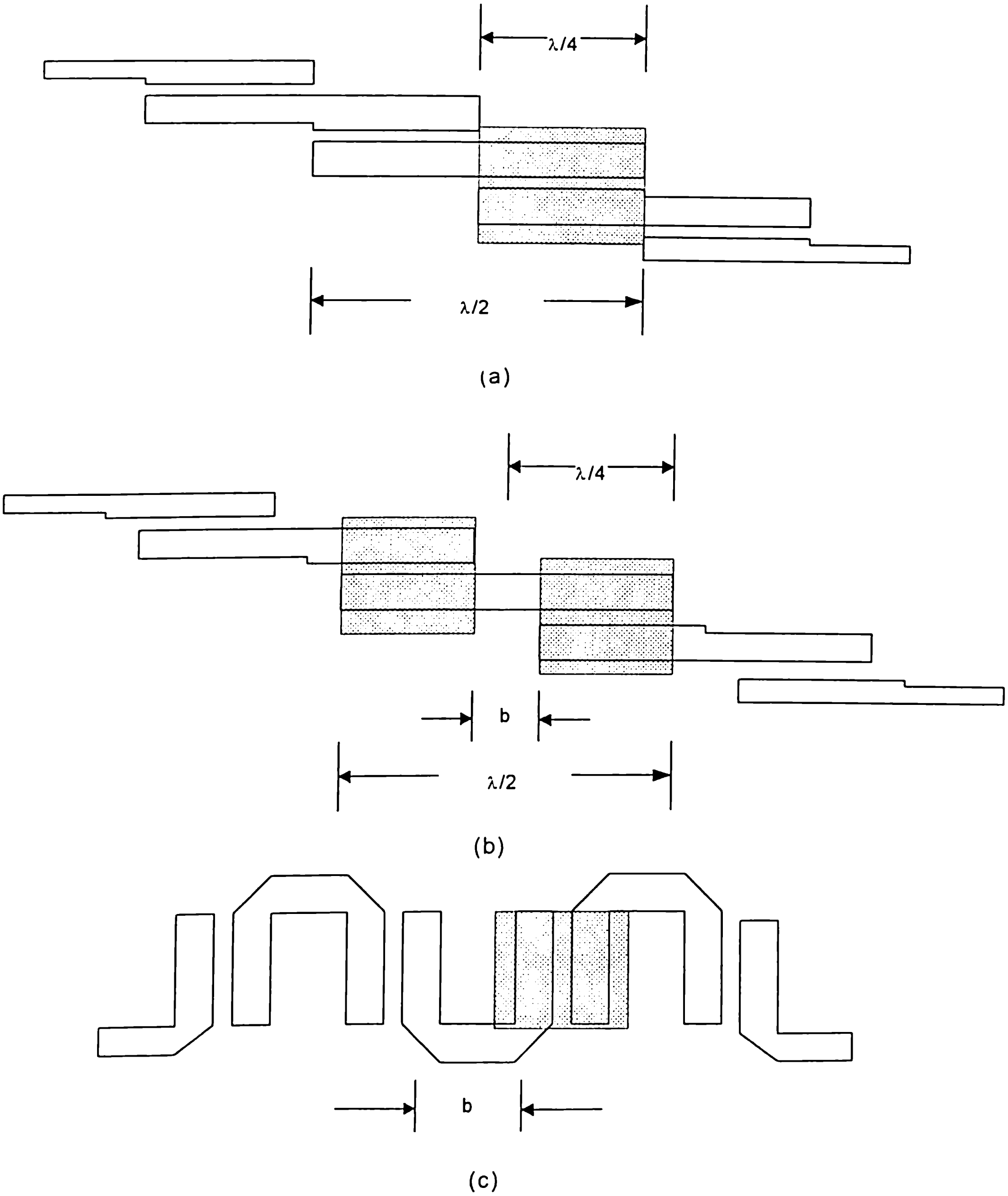


**Figure 1.** (**a**) Edge-coupled filter – the coupling resonators are λ/4 long (depicted by gray area); (**b**) A sliding factor, *b*, is introduced (notice the gray area where length < λ/4); (**c**) The bending of the λ/2 resonators is made at the area of the slide factor to produce the hairpin resonator structure.

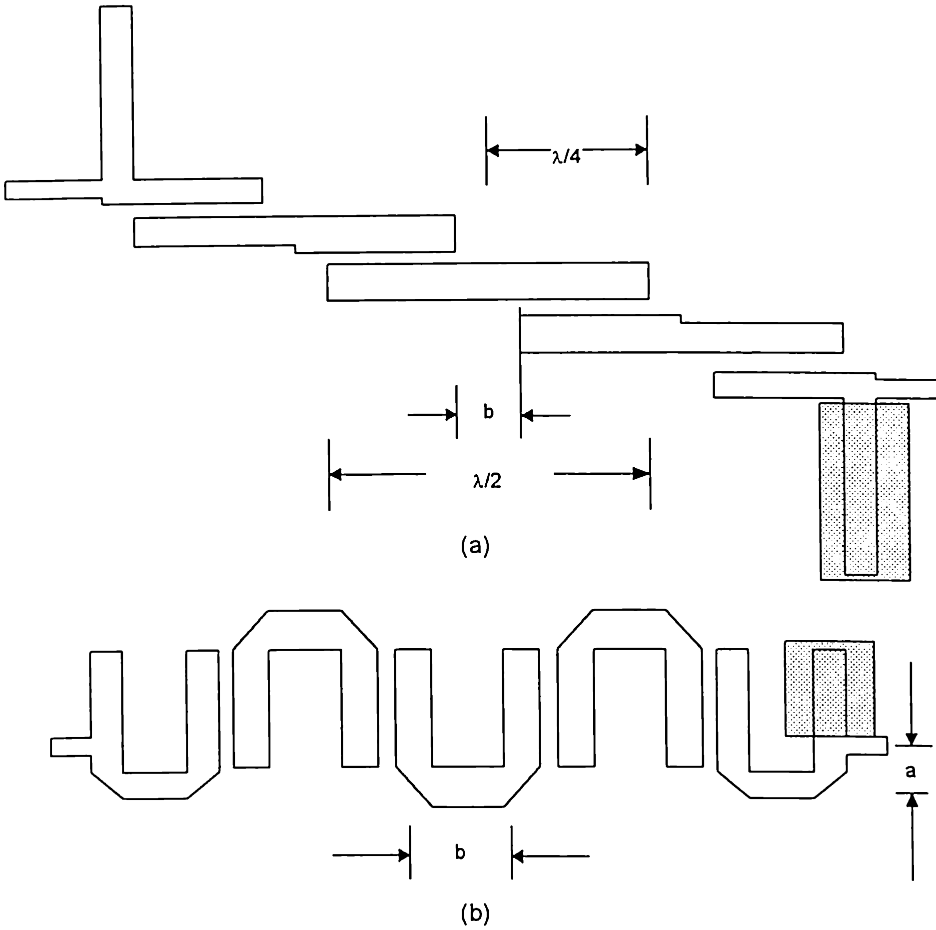


**Figure 2.** (a) Edge-coupled filter with matching stub (gray area); (b) The tapped hairpin resonator filter is formed by incorporating the matching stub into the end resonators. The tapping distance, *a*, is adjusted for optimal return loss.

An EDA software was initially used to optimize the design. The laminate parameters used in the software are listed in Table 1. These parameter values often lead to results that are several hundreds of megahertz away from the actual results. These values were based on the data sheets of Isola Laminate Systems (which is the manufacturer for the FR4 laminates used at ASTI) specifically its P.N. DURAVER E-CU#104 product, the analysis of microstrip line structures and [6].

**Table 1.**
Values for the laminate properties used in the EDA software.

| *Laminate Parameters* | *Value* |
|---|---|
| dielectric constant, $\varepsilon_r$ | *4.77* |
| laminate height, *H* | *59 mil* |
| metal thickness, *T* | *0.7 mil* |
| loss tangent, *Tan δ* | *0.008* |
| roughness, *r* | *0.075 mil* |
| relative permeability, *Mur* | *1* |
| conductor conductivity, *Cond* | *5.88E+07* |
| cover height, *hut* | *3.9E+34 mil* |

The changes in the frequency response of the filter were recorded during the network characterization of different filter samples. A specific structural dimension is altered in each sample to observe how it affects the overall filter response.

## III. DESIGN PATTERNS

To begin with the design, the resonator dimensions for an edge-coupled filter must be calculated using the steps given in [5]. Using an EDA program (LineCalc transmission line calculator [7]), the resonators' spacing, width and length were determined from the laminate parameters and the $Z_{oe}$ and $Z_{oo}$ impedance values calculated from [5]. An arbitrary slide factor was added to allow for resonator bending. The basic structure and the dimensional variables are shown in Fig. 3.

A $5^{th}$ order filter was used throughout the experiment regardless of bandwidth requirement since fewer orders would provide poor selectivity and more orders would make the design too large and introduce a higher insertion loss.

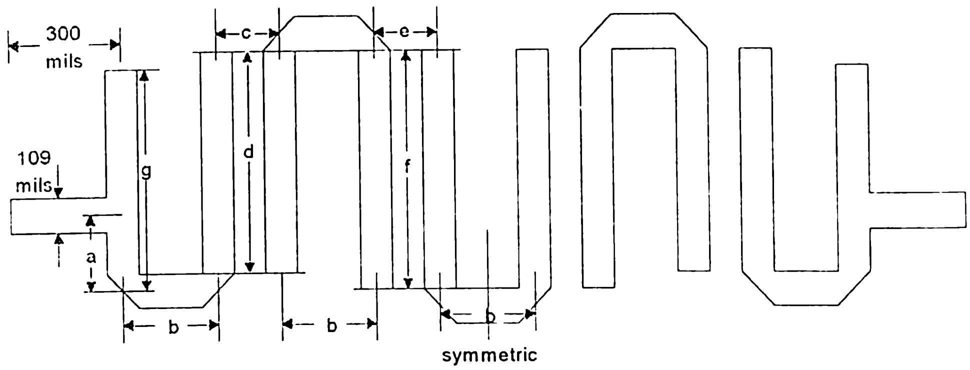


**Figure 3.** Tapped hairpin resonator filter dimensions. The structure is assumed to be symmetric so that $S_{11}=S_{22}$ and $S_{21}=S_{22}$ and that fewer variables could be used.

The dimension variables used in most of the experiments were:

| | |
|---|---|
| *a = 263 mils* | *e = 171 mils* |
| *b = 525 mils* | *f = 416 mils* |
| *c = 131 mils* | *g = 430 mils* |
| *d = 416 mils* | |

Note that the width of the 1st, 2nd and 3rd resonators according to the steps given in [5] should be 91 mils, 107.6 and 108 mils respectively. For design characterization simplification, the width of the structure within the filter was set at 108 mils while the tapping line is 109 mils (50 Ω).

### *3.1 Coupling Resonator Length, d*

The resonator length *d* was varied to see its effect on the overall frequency response. Designs were made for *d = 263 mils, d + 2 mm, d + 4 mm, d – 2 mm* and *d – 4 mm*. The center frequencies were recorded and plotted in Fig. 4. These curves can be used as reference when designing for the filter's center frequency. Note however that the curves are functions of the initial values of parameters *a* to *g*.

To illustrate the use of Fig. 4., let *x* be the center frequency desired. This frequency theoretically corresponds to coupling resonator length *y*. When the results of the actual test is off frequency *x* by $\Delta x$ (because the initial parameter values *a* to *g* are not the same), all we have to do is estimate from the curve the $\Delta y$ needed to arrive at the desired frequency. Although the center frequency will change when other dimensional variables are varied, the effect is not as severe as the effect of the resonator lengths. Succeeding iterations can easily compensate for such changes.

The equations of the curves in Fig. 4 are:

$$1^{st}\text{ resonator length } y = 247.3\, f_c^2 - 1835.3\, f_c + 4345.3 \qquad (1)$$

$$2^{nd},3^{rd}\text{ resonator length } y = 247.3\, f_c^2 - 1835.3\, f_c + 4331.15 \qquad (2)$$

### *3.2 Resonator Spacing, c and e*

The design approximation used in this experiment is $e = c + 1\ mm$ since more coupling is required in the input matching/resonating element. This assumption allows the manipulation of just one variable instead of two thereby making analysis easier. A more involved experiment can be done to determine the actual response of varying *c* and *e*.

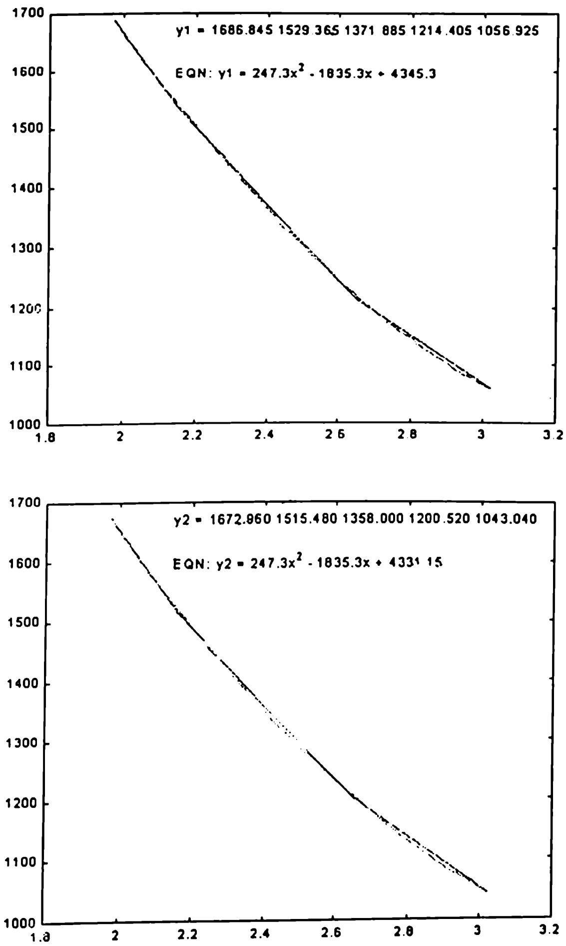


**Figure 4.** (a) Curve for the 1st resonator length. The y-axis gives the total length (i.e. $g + b + d$). (b) Curve for the 2st and 3rd resonator lengths. The y-axis gives the total length (i.e. $d + b + f$ or $2f + b$).

The resonator spacing affects the bandwidth of the filter. Varying the spacing changes the coupling between adjacent resonator elements. More signal power is transferred when a smaller spacing is used. The resulting filter response is a wider bandwidth as seen in Fig. 5. Unacceptable ripples are also present if the bandwidth is too wide.

The required resonance spacing for a desired bandwidth can be estimated from the curve in Fig. 5 in the same manner as the curves in Fig. 4 can be used to estimate the desired frequency. The bandwidth of the filter samples was measured between the points where the return loss becomes greater than or equal to 10dB.

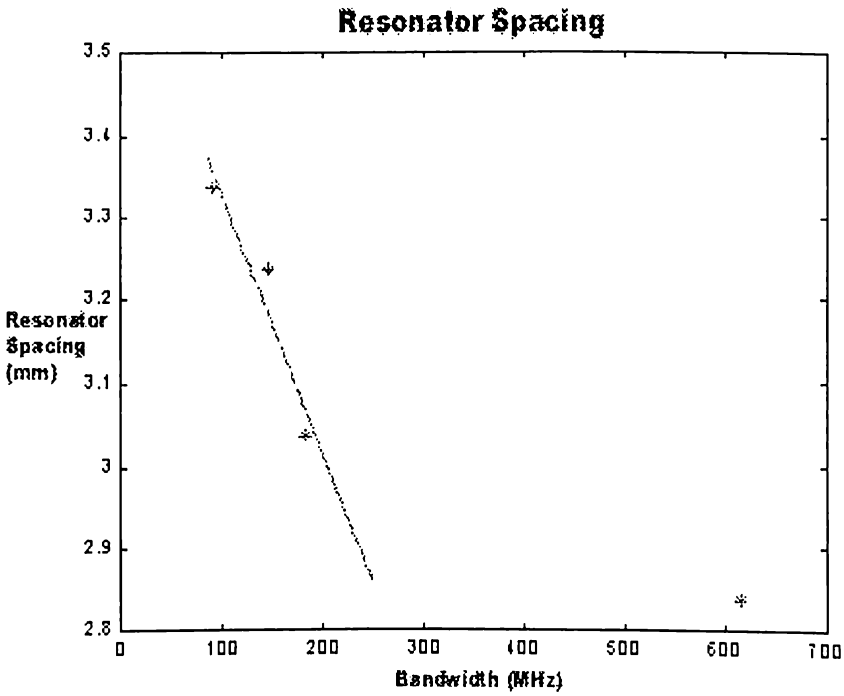


**Figure 5.** Resonator spacing vs. bandwidth.

Note that at $c = 2.84\ mm$, the bandwidth is very large (>600 MHz) and this point is not located in the linear curve. The hairpin filter is only intended for narrowband applications (about ≤ 10% of the center frequency). The adjacent resonator spacing will be so small (and thus making the filter difficult to fabricate) if the bandwidth is to be made larger. Undesirable ripples also appear making hairpin filters inappropriate for wideband applications.

### *3.3 Sliding Factor, b*

The sliding factor *b* affects the insertion loss and the possibility of reentrant frequencies. Since it is also part of the total resonator length, it also affects the center frequency.

When *b* is decreased, more of the resonator length form part of the coupling structure making energy transfer between resonator elements more efficient. This lowers the filter's insertion loss. However, decreasing *b* without holding the total resonator length constant also shifts the center frequency. It was observed though that the effect on the return loss is more evident if the total resonator length is not held constant. Note also that when the sliding factor is decreased too much, resonator self-coupling becomes significant. This was observed to produce unwanted reentrant frequencies near the desired frequency range as predicted in [4].

As the $S_{11}$ (return loss) did not significantly change when *d* was varied (and was therefore not reported in Sec. 3.1), it is expected that adjusting the resonator length after determining the optimum *b* to get the filter back to the desired frequency will not significantly affect the return loss. It is therefore suggested that experiments on decreasing b while keeping all other variables constant be done first to observe the varying $S_{11}$ response. Once the best $S_{11}$ response is determined, the resonator length can later be adjusted to obtain again the desired center frequency.

Note that although the tap distance *a* ultimately controls the return loss, the slide factor *b* maximizes the obtainable return loss.

### *3.4 Tap Distance, a*

Adjusting *a* is merely tuning the length of the filter's matching stub to obtain an impedance match. It was observed that the other properties of the filter like bandwidth and center frequency are almost unaffected by a varying tap distance. The tap distance is therefore the last variable being varied for optimal design.

## IV. PRACTICAL TECHNIQUES

To design a hairpin filter, first obtain the initial values of the dimensional variables by following the steps outlined in the beginning of Section 3 of this paper. Using these variables, construct a prototype filter test it, noting its center frequency and bandwidth.

Compare the center frequency results with the curves of Fig. 4. Adjust the total resonator lengths following the explanation given in Sec. 3.2. Using the same technique, determine the resonator spacing from the curve of Fig. 5 for the desired bandwidth. Create the next prototype and then test to verify. The resulting center frequency and bandwidth may not be exactly as planned but they should be close the desired values.

To determine the right sliding factor, first use an arbitrary sliding factor ($b = 525$ *mils* in the case of the experiment). Vary $b$ by a constant amount ($b = 525$ *mils* $- 2n$ *mm* where n = 1, 2, 3, ... in the case of the experiment) while holding all other variables constant. Do not mind the frequency shift. Choose the response with the best return loss. (The design iteration may take 3 prototype boards for this part.) This is approximately the correct sliding factor. Readjust $d, f$ and $g$ using Fig. 4 in order to recover the desired center frequency. The change in the insertion loss and return loss will just be slight.

Once the desired resonator lengths, resonator spacing and sliding factor have been determined, design a few more boards with varying tap distance to get the optimal design. Do not go beyond half of the resonator length $g$ as the response would not be good. By rule of thumb, the desired bandwidth should have a return loss of 10 dB at least.

The expected total number of boards to be used in this practical design approach should be about 8-10 boards. This is a lot less than the 32 boards used in this research to produce an optimal 2.32-2.36 GHz band pass filter. (This filter is being used in one of the transmit-receive pair of an E1 wireless transmission system under the Digital Microwave Radio project of the Advanced Science and Technology Institute. The filter must be able to operate between 2.32 - 2.36 GHz. The filter response at 2.45 GHz should be about 15 dB below the average insertion loss between 2.32-2.36 GHz thereby providing an additional 15 dB LO rejection to the mixer. The insertion loss at the other transmit-receive pair – 2.54-2.58 GHz – must be as high as possible.)

The optimal filter prototype is shown in Fig. 6 and its frequency response in Fig. 7. The final filter has a 6 dB insertion loss, a very good result considering the fact that small and steep bandwidths generally suffer from high insertion loss and that above 2 GHz, the FR4 laminate is very lossy.

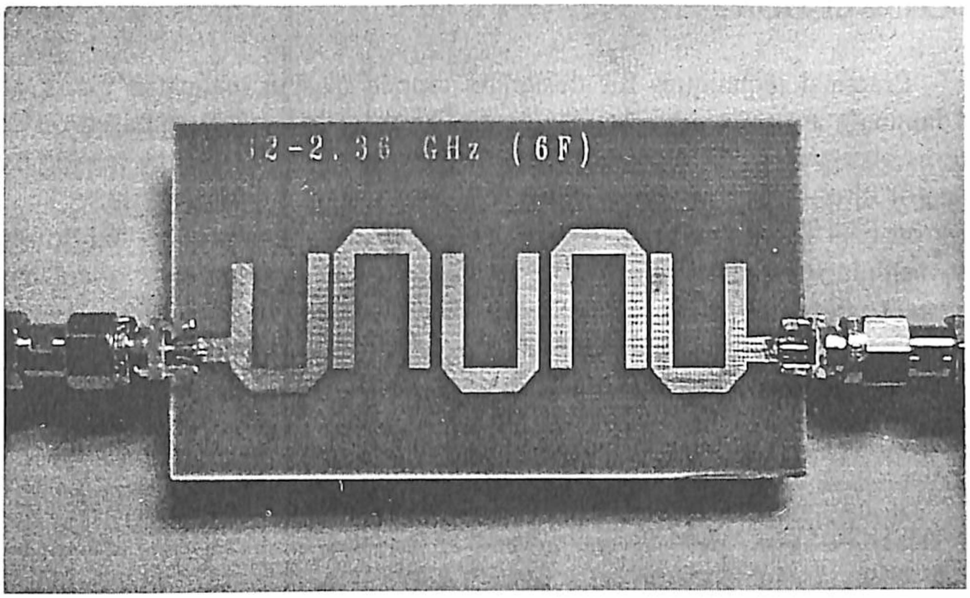


**Figure 6.** A 2.32-2.36 GHz tapped hairpin resonator filter built on an FR4 laminate. The connectors used are SMA connectors.

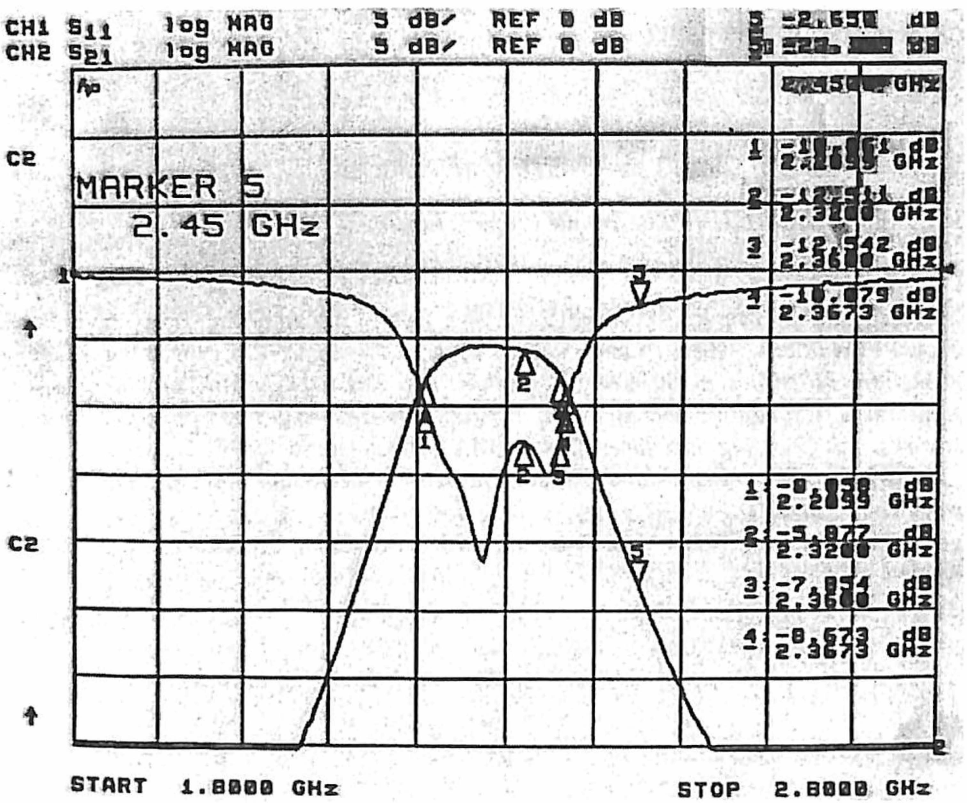


**Figure 7.** $S_{21}$ (bell-shaped) and $S_{11}$ curves of the filter as observed from an HP8753C Vector Network Analyzer. The vertical scale is 5dB/div and the frequency range is from 1.8 GHz to 2.8 GHz. The insertion loss is about 6 dB while the return loss is greater than 10 dB.

## V. CONCLUSION

Practical techniques for designing tapped hairpin resonator filters on the FR4 laminate are presented in this paper. These techniques were based on design patterns observed while varying several dimensional variables of the tapped hairpin resonator filter. This motivation is due to the difficulty of designing these filters on FR4 laminates using EDAs. Several patterns were observed during the filter sample characterization. The resonator length was shown to have a significant effect on the center frequency. The resonator spacing generally controls the bandwidth. Choosing the right slide factor increases the potential of the filter for a good match and low loss. The tap distance ultimately matches the filter to the rest of the circuitry.

## VI. ACKNOWLEDGMENT

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