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Design Of Triple Band Pass Filters Using Three-Coupled Finline And Metamaterials

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Abstract : This paper presents a new concept of a triple band pass filters with three parallel coupled finline structure and Metamaterials. The design of third order bandpass filter, having center frequency of 10 GHz with fractional bandwidth of 20 % is simulated in High Frequency Structure Simulator (HFSS). A design graph is presented here for symmetric three unilateral finline structures to design wide bandpass filter parameters like length (l), width (w) and spacing(s) between finlines. This wide band pass filter is converted in to triple band pass filter with Metamaterials i.e. (Split Ring Resonators) on the other side of this filter structure substrate is designed and simulated with HFSS.

Keywords : Unilateral finline, Multi conductor lines, Wide bandpass filter, Metamaterials, Split Ring Resonators (SRR), Triple band pass filters.

I. INTRODUCTION

THE finline is transmission line viz. metallic strips etched on substrate embodied in the trunk of the standard rectangular wave guiding structure which is increasingly used as millimeter wave component due to various advantages such as reducing size, weight and cost. At millimeter wave frequency the finline filter has been implemented in [1]-[6] which are mostly based on ladder/cascaded shape. This paper presents a Chebyshev filter of order 3 with fractional bandwidth 20% has been designed on RT-duroid 5880™ substrate using unilateral three coupled finline. The advantage of this filter is low loss and wider bandwidth over the ladder/cascaded type filter.

Metamaterials is a very interesting concept for converting this wide band pass filter into more band pass filters for effective utilization this band width. Metamaterials in microwave field have been an object of great interest in recent past years. The hypothesis of V.G. Veselago in 1968, [8] that materials with simultaneous negative effective permittivity $\varepsilon_{eff}(\omega)$ and negative effective permeability μ_{eff} (ω) have unusual reversed electromagnetic wave propagation phenomena. Naturally occurring materials universally have a positive permeability and thus a Left-Handed Material (LHM), while not ruled out by fundamental considerations, seemed unlikely to be practical. However, in 1999, Pendry et al, [9] introduced several configurations of conducting scattering elements displaying a magnetic response to an applied electromagnetic field when grouped in to an interacting periodic array. SRR creates negative effective permeability μ_{eff} (ω) over a particular frequency region and wire elements to produce negative effective permittivity $\varepsilon_{eff}(\omega)$ in an intersecting frequency region.

In this paper numerical procedure based on the full wave modal analysis is formulated to compute all the frequency-dependent normal mode parameters for symmetric unilateral finlines. A bandpass filter is designed using full wave modal analysis is presented in Section.

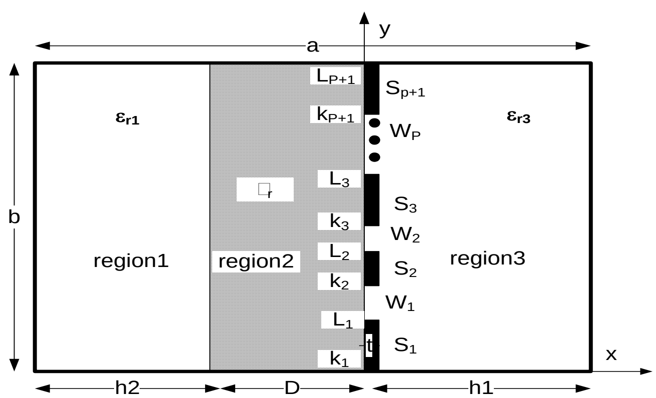

Fig. 1: Cross-section view of multiple edge coupled unilateral fin-line on isotropic substrate.

Analysis of Three Coupled Unilateral Finlines

The dispersion characteristics of multiple coupled Fin lines on isotropic substrate are evaluated by using full wave modal analysis. In the modal analysis all the field components are constructed in terms of x-components of electric and magnetic fields in each region, which are expanded in terms of modal fields with unknown coefficients as given below.

The dispersion characteristics of unilateral fin-line have been computed by choosing the following slot field distributions.

$$\begin{bmatrix} E_t \\ H_t \end{bmatrix} = \frac{1}{\left(k_o^2 \varepsilon_r - k_x^2\right)} \begin{bmatrix} \partial/\partial x & j\omega\mu_o x \\ -j\omega\varepsilon_o\varepsilon_r x & \partial/\partial x \end{bmatrix} \begin{bmatrix} \nabla_t E_x \\ \nabla_t H_x \end{bmatrix} \quad (1)$$

and the wave equation can be written as

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$$\left[\frac{\partial^2}{\partial x^2}+\frac{\partial^2}{\partial y^2}+\left(k_o^2\varepsilon_r-\beta^2\right)\right]\begin{bmatrix}E_x\\H_x\end{bmatrix}=0 \qquad (2)$$

Here side walls are considered to be electric walls. Solution of equation (2) in three regions are

$$E_x^{(1)}=\sum_{n=1}^{\infty}A_{n1}Cos\left[\Gamma_{n1}(x-h_1)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$H_x^{(1)}=\sum_{n=0}^{\infty}B_{n1}Sin\left[\Gamma_{n1}(x-h_1)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$E_x^{(2)}=\sum_{n=1}^{\infty}\left[A_{n2}Sin(\Gamma_{n2}x)+A'_{n2}Cos(\Gamma_{n2}x)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$H_x^{(2)}=\sum_{n=0}^{\infty}\left[B_{n2}Cos(\Gamma_{n2}x)+B'_{n2}Sin(\Gamma_{n2}x)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$E_x^{(3)}=\sum_{n=1}^{\infty}A_{n3}Cos\left[\Gamma_{n3}(x+d+h2)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$H_x^{(3)}=\sum_{n=0}^{\infty}B_{n3}Sin\left[\Gamma_{n3}(x+d+h2)\right]Cos(\alpha_n y)e^{-j\beta z}$$

Where

$$\Gamma_{n1}=\sqrt{k_0^2-\alpha_n^2-\beta^2}$$

$$\Gamma_{n2}=\sqrt{k_0^2\varepsilon_{r2}-\alpha_n^2-\beta^2}$$

$$\Gamma_{n3}=\sqrt{k_0^2\varepsilon_{r3}-\alpha_n^2-\beta^2}$$

$$\alpha_n=\frac{n\pi}{b}$$

In the field equations, the boundary conditions at x = h1, (-h2+d) and y= ± b/2 have been incorporated. Other field components can be derived as

$$E_y^{(1)}=\sum_{n=0}^{\infty}S_{n1}Sin\left[\Gamma_{n1}(x-h_1)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$E_y^{(2)}=\sum_{n=0}^{\infty}\left[S_{n2}Cos(\Gamma_{n2}x)+S'_{n2}Sin(\Gamma_{n2}x)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$E_y^{(3)}=\sum_{n=0}^{\infty}S_{n3}Sin\left[\Gamma_{n3}(x+d+h2)\right]Cos(\alpha_n y)e^{-j\beta z}$$

Where,

$$S_{n1}=\frac{1}{\alpha_n^2+\beta^2}\left[-\alpha_n\Gamma_{n1}A_{n1}-\omega\mu_o\beta B_{n1}\right]$$

$$S_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[\alpha_n\Gamma_{n2}A_{n2}-\omega\mu_o\beta B_{n2}\right]$$

$$S'_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[-\alpha_n\Gamma_{n2}A'_{n2}-\omega\mu_o\beta B'_{n2}\right]$$

$$S_{n3}=\frac{1}{\alpha_n^2+\beta^2}\left[-\alpha_n\Gamma_{n3}A_{n3}-\omega\mu_o\beta B_{n3}\right]$$

$$E_z^{(1)}=\sum_{n=1}^{\infty}C_{n1}Sin\left[\Gamma_{n1}(x-h_1)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$E_z^{(2)}=\sum_{n=1}^{\infty}\left[C_{n2}Cos(\Gamma_{n2}x)+C'_{n2}Sin(\Gamma_{n2}x)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$E_z^{(3)}=\sum_{n=1}^{\infty}C_{n3}Sin\left[\Gamma_{n3}(x+d+h2)\right]Sin(\alpha_n y)e^{-j\beta z}$$

Where

$$C_{n1}=\frac{1}{\alpha_n^2+\beta^2}\left[j\beta\Gamma_{n1}A_{n1}-j\omega\mu_o\alpha_n B_{n1}\right]$$

$$C_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[-j\beta\Gamma_{n2}A_{n2}-j\omega\mu_o\alpha_n B_{n2}\right]$$

$$C'_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[j\beta\Gamma_{n2}A'_{n2}-j\omega\mu_o\alpha_n B'_{n2}\right]$$

$$C_{n3}=\frac{1}{\alpha_n^2+\beta^2}\left[j\beta\Gamma_{n3}A_{n3}-j\omega\mu_o\alpha_n B_{n3}\right]$$

$$H_y^{(1)}=\sum_{n=1}^{\infty}M_{n1}Cos\left[\Gamma_{n1}(x-h_1)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$H_y^{(2)}=\sum_{n=1}^{\infty}\left[M_{n2}Sin(\Gamma_{n2}x)+M'_{n2}Cos(\Gamma_{n2}x)\right]Sin(\alpha_n y)e^{-j\beta z}$$

$$H_y^{(3)}=\sum_{n=1}^{\infty}M_{n3}Cos\left[\Gamma_{n3}(x+d+h2)\right]Sin(\alpha_n y)e^{-j\beta z}$$

Where

$$M_{n1}=\frac{1}{\alpha_n^2+\beta^2}\left[\omega\varepsilon_o\beta A_{n1}-\alpha_n\Gamma_{n1}B_{n1}\right]$$

$$M_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[\omega\varepsilon_o\varepsilon_{r2}\beta A_{n2}+\alpha_n\Gamma_{n2}B_{n2}\right]$$

$$M'_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[\omega\varepsilon_o\varepsilon_{r2}\beta A'_{n2}-\alpha_n\Gamma_{n2}B'_{n2}\right]$$

$$M_{n3}=\frac{1}{\alpha_n^2+\beta^2}\left[\omega\varepsilon_o\varepsilon_{r3}\beta A_{n3}-\alpha_n\Gamma_{n3}B_{n3}\right]$$

$$H_z^{(1)}=\sum_{n=0}^{\infty}D_{n1}Cos\left[\Gamma_{n1}(x-h_1)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$H_z^{(2)}=\sum_{n=0}^{\infty}\left[D_{n2}Sin(\Gamma_{n2}x)+D'_{n2}Cos(\Gamma_{n2}x)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$H_z^{(3)}=\sum_{n=0}^{\infty}D_{n3}Cos\left[\Gamma_{n3}(x+d+h2)\right]Cos(\alpha_n y)e^{-j\beta z}$$

$$D_{n1}=\frac{1}{\alpha_n^2+\beta^2}\left[-j\omega\varepsilon_o\alpha_n A_{n1}-j\beta\Gamma_{n1}B_{n1}\right]$$

$$D_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[-j\omega\varepsilon_o\varepsilon_{r2}\alpha_n A_{n2}+j\beta\Gamma_{n2}B_{n2}\right]$$

$$D'_{n2}=\frac{1}{\alpha_n^2+\beta^2}\left[-j\omega\varepsilon_o\varepsilon_{r2}\alpha_n A'_{n2}-j\beta\Gamma_{n2}B'_{n2}\right]$$

$$D_{n3}=\frac{1}{\alpha_n^2+\beta^2}\left[-j\omega\varepsilon_o\varepsilon_{r3}\alpha_n A_{n3}-j\beta\Gamma_{n3}B_{n3}\right]$$

We now apply the continuity conditions at x=0.They are

$$H_x^{(1)} = H_x^{(2)}$$
$$E_y^{(1)} = E_y^{(2)}$$

The continuity condition at x=-d are given by

$$E_y^{(3)} = E_y^{(2)}$$
$$E_z^{(3)} = E_z^{(2)}$$
$$H_y^{(3)} = H_y^{(2)}$$
$$H_z^{(3)} = H_z^{(2)}$$

$$\begin{bmatrix} A_{n3} \\ B_{n3} \end{bmatrix} = \begin{bmatrix} -\frac{\varepsilon_{r2}}{\varepsilon_{r3}} & 0 & \frac{\varepsilon_{r2}}{\varepsilon_{r3}} & 0 \\ 0 & \frac{\Gamma_{n2}}{\Gamma_{n3}} & 0 & \frac{\Gamma_{n2}}{\Gamma_{n3}} \end{bmatrix} \begin{bmatrix} A_{n2} \frac{Cos(\Gamma_{n2} d)}{Sin(\Gamma_{n3} h_2)} \\ B_{n2} \frac{Cos(\Gamma_{n2} d)}{Sin(\Gamma_{n3} h_2)} \\ A'_{n2} \frac{Sin(\Gamma_{n2} d)}{Sin(\Gamma_{n3} h_2)} \\ B'_{n2} \frac{Sin(\Gamma_{n2} d)}{Sin(\Gamma_{n3} h_2)} \end{bmatrix}$$

Equating the values of A_{n3} & B_{n3} we get

$$A'_{n2} = A_{n2} \left[\frac{\varepsilon_{r2}\Gamma_{n3} Sin(\Gamma_{n2} d) Sin(\Gamma_{n3} h_2) - \varepsilon_{r3}\Gamma_{n2} Cos(\Gamma_{n2} d) Cos(\Gamma_{n3} h_2)}{\varepsilon_{r2}\Gamma_{n3} Cos(\Gamma_{n2} d) Sin(\Gamma_{n3} h_2) + \varepsilon_{r3}\Gamma_{n2} Sin(\Gamma_{n2} d) Cos(\Gamma_{n3} h_2)} \right]$$

and

$$B'_{n2} = B_{n2} \left[\frac{\Gamma_{n3} Cos(\Gamma_{n2} d) Cos(\Gamma_{n3} h_2) - \Gamma_{n2} Sin(\Gamma_{n2} d) Sin(\Gamma_{n3} h_2)}{\Gamma_{n3} Sin(\Gamma_{n2} d) Cos(\Gamma_{n3} h_2) + \Gamma_{n2} Cos(\Gamma_{n2} d) Sin(\Gamma_{n3} h_2)} \right]$$

Hence all the field coefficients can be represented in terms of A_{n1} and B_{n1}.

II. FILTER DESIGN

The wide band pass filter is designed with three parallel-coupled unilateral finlines approximately quarter wavelength long. In this paper multi resonators are cascaded to achieve high rejections. The six port impedance matrix parameters for a section of three coupled finlines of length *l* are found from mode characteristic impedances, phase velocities and voltage ratios.

This three-coupled finline structure supports three dominant modes as OE, EE, and OO, which correspond to 1, 2 and 3, respectively. Each mode has its own modal phase constant, eigen voltage vector and characteristic impedance. The eigen voltage matrix for symmetrical three line which have equal fin-width and spacing are given by

$$[M_V] = \begin{bmatrix} 1 & 1 & 1 \\ m_1 & 0 & m_3 \\ 1 & -1 & 1 \end{bmatrix} \quad (3)$$

each vector of $[M_V]$ is the eigen voltage vector of the matrix product [L] [C]. The matrix $[M_V]$ can be used to derive the relation between port voltages and port currents.

$$\begin{bmatrix} V_A \\ V_B \end{bmatrix} = \begin{bmatrix} Z_A & Z_B \\ Z_B & Z_A \end{bmatrix} \begin{bmatrix} I_A \\ I_B \end{bmatrix} \quad (4)$$

Where

$$[V_A] = [V_1, V_2, V_3]^T, [V_B] = [V_4, V_5, V_6]^T,$$
$$[I_A] = [I_1, I_2, I_3]^T, [I_B] = [I_4, I_5, I_6]^T$$

And the impedance matrix $[Z_A]$ and $[Z_B]$ can be derived as

$$[Z_A] = [M_V] \ diag[-jZ_{mi} \cot\theta_i][M_V]^T \quad (5)$$

$$[Z_B] = [M_V] \ diag[-jZ_{mi} \csc\theta_i][M_V]^T \quad (6)$$

Now $\theta_i = \beta_i l$ with β_i is the phase constant of the i^{th} mode, *l* the length of the coupled section, and Z_{mi} given by

$$Z_{mi} = \frac{Z_{oi}}{m_i^2 + 2} \quad (7)$$

Where Z_{oi} is the characteristic impedance of i^{th} mode.
Eqn (7), (8), (9) are derived from eqn. (6).

$$m_1 Z_{m1} - m_3 Z_{m3} = J Z_A Z_B \quad (8)$$
$$m_1^2 Z_{m1} - m_3^2 Z_{m3} = Z_A (J^2 Z_A Z_B + 1) \quad (9)$$
$$Z_{m1} + Z_{m3} = Z_B (J^2 Z_A Z_B + 1) \quad (10)$$

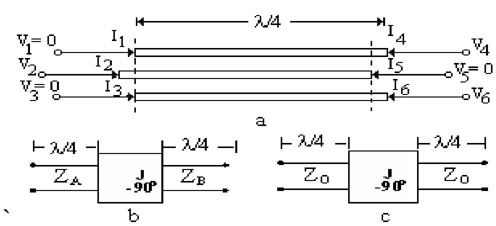

Fig.2. Reduction of a coupled three-finline section to a two-port network (a) Coupled three-line section as a six-port network (b) Equivalent admittance inverter (c) Further approximated admittance inverter.

After simplifying eqn. (8), (9) & (10) we obtain the below equations

$$m_1 Z_{m1} \approx \left[\frac{2+\mu}{2\mu} \right]^2 (Z_0/2)(J^2 Z_0^2 + J Z_0 + 1) \quad (11)$$

$$m_3 Z_{m3} \approx \left[\frac{2+\mu}{2\mu} \right]^2 (Z_0/2)(J^2 Z_0^2 - J Z_0 + 1) \quad (12)$$

where

$$\mu = \frac{\sqrt{2\left[2(Z_{ee} - Z_{oo})^2 - Z_{oe}(Z_{oe} - Z_{ee} - Z_{oo}) - Z_{ee} Z_{oo}\right]}}{2Z_{ee} - Z_{oe} - Z_{oo}}$$

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The value of JZ_O for each admittance inverter can be determined from the values of lumped circuit elements of the low pass prototype [16].

$$J_1 = \frac{1}{Z_0}\sqrt{\frac{\pi\Delta}{2g_1}} \tag{13}$$

$$J_n = \frac{1}{Z_0}\sqrt{\frac{\pi\Delta}{2g_{n-1}g_n}},\ for\ n = 2,3.....N, \tag{14}$$

$$J_{N+1} = \frac{1}{Z_0}\sqrt{\frac{\pi\Delta}{2g_N g_{N+1}}} \tag{15}$$

where $\Delta = \frac{\omega_2 - \omega_1}{\omega_0}$

For N = 3, the values of g_1 to g_{N+1} are given below.
g_1 = 1.5963, g_2 = 1.0967, g_3 = 1.5963, g_4 = 1.0000.

Once JZ_O is known the values of m_1Z_{m1} and m_3Z_{m3} for each coupled section can be known. The values of m_1Z_{m1} and m_3Z_{m3} for section I are 82.02 Ω and 64.69 Ω respectively and for the section II are 37.65 Ω and 40.94 Ω respectively. The designed data from the above is calculated and finally the filter is optimized. Both the data are shown in Table I.

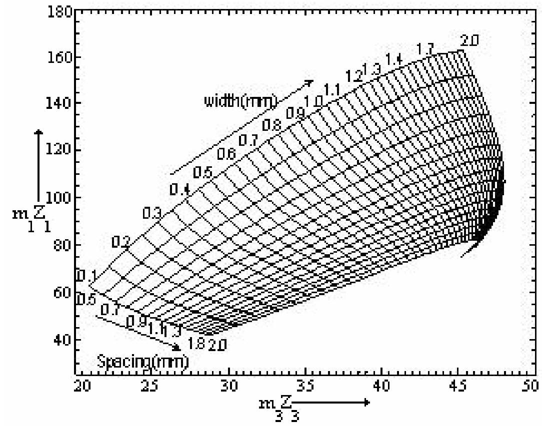
TABLE I

Dimentions	Designed Data		Optimized Data	
	Section 1	Section 2	Section 1	Section 2
W	0.650	0.68	0.750	0.750
S	1.320	2.370	1.400	2.400
L	9.592	10.164	12.970	13.413
G	1.975	2.033	1.076	1.633

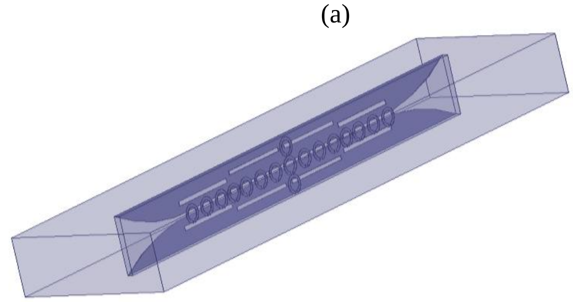
All dimensions are in mm.

The design graph in Fig. 3(a), for ε_r=2.2 is used to determine the line width and line spacing of three coupled unilateral finlines at 10 GHz. Due to symmetry of filter only the width, spacing and length of two sections has been mentioned here. The simulated three dimension model of three finline wide bandpass filter structure is shown in Fig. 3(b). This parallel arrangement gives relatively large coupling for a given spacing between resonators, and thus this filter structure is particularly convenient for large bandwidths as compared to the other structures [10].

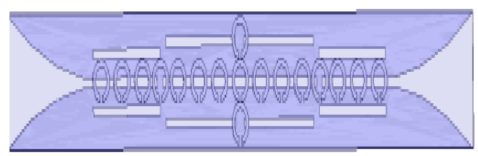
Now Split Ring Resonator (SRR) concept has been introduced on other side of substrate of this wide band pass filter. These SRRs creates negative effective permeability $\mu_{\mathbf{eff}}$ (ω) at desired frequency. It creates a notch at resonance frequency. This is the useful characteristics for converting the wideband pass filter in to triple bandpass filters. The notch can be controlled by number of SRRs and placing. This is a new technique in finline technology to develop a triple bandpass filters. Split Ring Resonator (SRR) can be seen as a corresponding LC circuit. The response this triple bandpass filter is shown in fig.4.



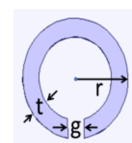
(a)



(b)



(c)



(d)

Fig.3. (a) The bandpass filter design graph for a symmetric three unilateral finline structure. [Substrate dielectric constant εr=2.2, substrate thickness (D) = 0.8 mm, frequency =10 GHz]. (b)Simulated structure of three coupled fin-line filter with X-band wave housing. (c) PCB layout of the wide bandpass filter (three coupled finline of order 3). (d) SRR dimensions: r (radius of ring) = 1.5mm, t (thickness of ring) = 0.4mm and g (gap of ring) = 0.4mm.

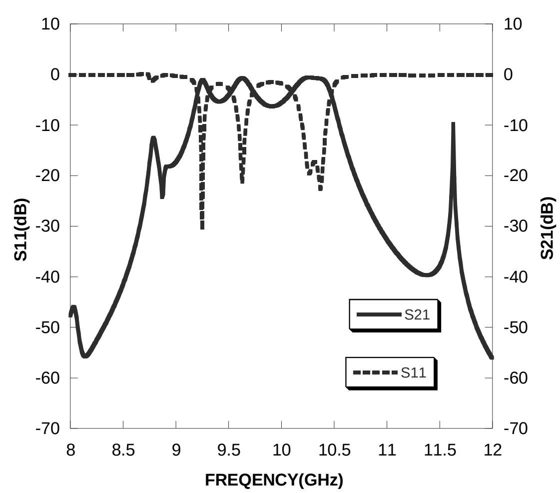


Fig.4. Response of triple-band pass three finline filter.

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Some of the important observations are given as: (1) Large band separation and sharp response is possible with more number of SRR and their placing of the structure. (2) Again triple bandpass is tradeoff between band separation and return losses. (3) At center of the structure, two extra SRRs are included at top and bottom. These SRRs are useful for S11is less than -20dB and sharp response, without these two SRRs the magnitude of third band S11 is nearly -10dB.

III. CONCLUSION

Triple bandpass filters play an important role in the modern transceivers. The most important advantages are, these can be widely used in millimeter wave applications such as automotive radar and Radio-Frequency Identification (RFID) systems, transceivers for portable wireless communication devices. Modern communication transceivers require high performance microwave filters with low insertion loss, high frequency selectivity and small group delay variations. Most of the above said parameters are obtained with this proposed triple bandpass filter with three coupled finlines with metamaterials.

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