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INTERFERENCE CANCELLATION SYSTEM OF A MICROWAVE RESONATOR FILTER FOR 4G AND 5G APPLICATIONS.

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ABSTRACT

This paper aimed at designing a microwave resonator filter for use in 4G and 5G applications. A band resonator filter was designed and implemented using microstrip technology. Poor performance of a microstrip microwave resonator filter is due to spurious harmonics and unwanted frequencies which are always present and cause interferences of signals hindering effective communication. The coupling matrix was first obtained using polynomial functions of transmission and reflection coefficients. A cost function for optimization was developed and particle swarm optimization techniques were used to optimize the coupling matrix. The transversal matrix was developed and converted to the folded matrix for practical filter implementation. A bandpass filter was designed and simulated using high-frequency structure simulator software. Unwanted frequencies are eliminated by employing a novel square open loop resonator structure. The filter was designed to specification and the results show that unwanted frequencies were eliminated corresponding to the frequency of the outer resonator and therefore performs better than comparable filter.

Keywords: Harmonics, microstrip technology, square open-loop resonator, triple band resonator filter, high-frequency structure simulator.

INTRODUCTION

Microwave filters are present in most communication systems and are key components of modern communication systems and networks. The microwave communication system has been undergoing serial developments for more than four decades which resulted in overcrowding of the lower frequency spectrum as the development of mobile communication systems, radar technology, and broadcasting continue to increase in sophistication. This places a bigger demand on the higher frequency spectrum of the electromagnetic spectrum as microwave filters are used for processing this high-frequency signal (Xia, 2015., Lee, 2009). Harmonics are multiples integral of frequencies which cause distortion and interferences in microwave filters. They are present at first harmonic which is two times fundamental frequencies $(2f_0)$ in loop resonator filters (Atallah, 2010). Several researchers have done work involving different approach of reducing spurious harmonics. Sobhan, R. et al. (2023) developed use of open and short stub and lumped components inorder to reduce harmonics present in power divider. The researcher finds out that use of coupled lines structure improves its performances in terms of size reduction and filtering capabilities. This compares with the traditional method of harmonic reductions involving the use of open and short stubs, capacitors and inductors component and coupled line structure (Sobhan et al., 2023). Similarly, Phromloungsri, R. et al.(2005) investigated inductors coupled to its ports to improve directivity and hence reduce harmonics in bandpass filter.

Muhammed, R. (2017) examined a quasi-ellipticalfunction bandpass filter having its input and output feedlines connected with stubs of inductive composition to reduce the effect of harmonics. The bandpass filter was designed at a centre frequency of 3.2GHz with harmonics suppressed up to about $3f_0$.

Moreover Jitha, B. (2010) investigated a microstripbandpass filter designed at a centre frequency of 6GHz. The researcher connected capacitive stubs to both input and output feedlines and harmonics remains at frequency below 6GHz.

On the other hand, King et al (2020) examined the suppression of harmonics in a waveguide bandpass filter at afrequency of 30GHz using impedance K-inverters. This widen the stopbandwith lower order harmonics being suppressed. Higher order harmonics still pose a significant threat.

Furthermore, Falih, M.A. et al. (2021) examined stub loaded multiple mode resonator designed at a centre frequency of about 3.95GHz. The addition of stubs led to the reduction of spurious harmonics but the method employed is complex.

Mark, H. (2016) investigated a reconfigurable bandstop resonator filter covering 75-110GHz frequencies ranges with higher order mode harmonics present.

Salah, I.Y. et al. (2023) examined planarbandpass resonator filter harmonics reduction using rat-race coupler. The filter was designed to operate at a centre frequency of 950MHz with harmonics reduced.

In addition, Abdel, A.S. and Ibrahim, E. (2024) investigated the removal spurious harmonics in a stepped impedance resonator filter using quarter wave stubs coupled to the filter. This widens the stopband thereby reducing spurious harmonics but higher order mode harmonics persists.

Futhermore, Somchat, and Ravee, P. (2018) developed a method of removing harmonics by introducing bandstop filters at its feedlines or port which successfully suppresses the harmonics present. Kopang, J. (2018) developed an interference cancellation system using amplitude and phase cancellation between two connected branches of a diplexer. This method becomes more complex if the resonator order is high. Higher resonator order improves the selectivity of the filter.

This research removes not only harmonics but any given frequencies of interest. The result shows that all unwanted frequencies are successfully removed. A better alternative approach to eliminating unwanted frequencies and harmonics with less complexity or simple configuration is developed using this novel approach.

MATERIALS AND METHOD

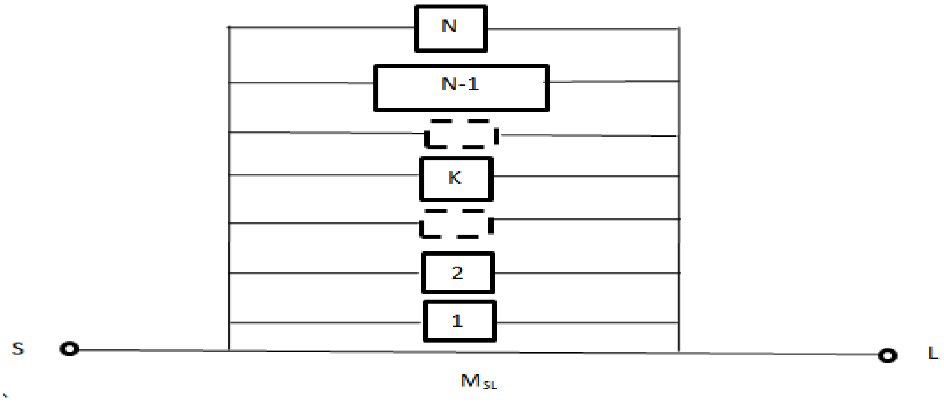
The method adopted involves:

(i) Design of microwave resonator filter

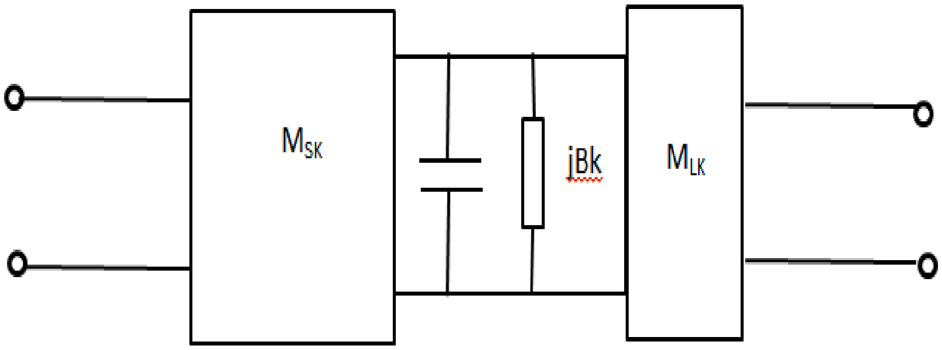
(ii) Computer-aided design using High-frequency structure simulator software

Microwave filter design.

The transversal matrix is made up of lowpass filters connected in parallel between source and load as shown in Figure 1b



(a) N- resonator transversal network including direct source load coupling M_{SL}



(b)

Figure 1: Transversal model (a) N resonator Transversal model (b) Lowpass equivalent circuit of the transversal model (Iman, 2013)

However, the synthesis of the N+2 transversal coupling matrix takes a similar approach as to the NxN coupling matrix. Y-parameters are derived from s-parameter transfer and reflection polynomials to obtain equation (1)

In circuit form, the admittance matrix $[Y]$ of the transversal matrix becomes

$$Y_N = j\begin{bmatrix} 0 & M_{SL} \\ M_{SL} & 0 \end{bmatrix} + \sum_{K=1}^{N} \frac{1}{sC_K - jB_k}\begin{bmatrix} M_{SK}^2 & M_{SK}M_{LK} \\ M_{SK}M_{LK} & M_{LK}^2 \end{bmatrix} (1)$$

And also y-parameters derived from the transversal network of Figure 1 yield

$$Y_N = \begin{bmatrix} y_{11k(s)} & y_{12k(s)} \\ y_{21k(s)} & y_{22k(s)} \end{bmatrix}$$

$$= j\begin{bmatrix} 0 & K_0 \\ K_0 & 0 \end{bmatrix} + \sum_{K=1}^{N} \frac{1}{s - j\lambda_k}\begin{bmatrix} r_{11k} & r_{12k} \\ r_{21k} & r_{22k} \end{bmatrix} (2)$$

By equating equation (1) and equation (2), elements of the transversal matrix can be obtained as:

$M_{SL} = K_0$

$C_K = 1, \quad B_K = -\lambda_K$

$$M_{LK}^2 = r_{22k} \text{ i.e } M_{LK} = \sqrt{r_{22k}} = T_{NK} \qquad (3)$$

$$M_{SK}M_{LK} = r_{21k} \text{i.e } M_{SK}^2 = \frac{r_{21k}}{\sqrt{r_{22k}}} = T_{1K}$$

The capacitors C_k are all taken as unity, susceptance $B_K = -\lambda_{k,}$ represent self-coupling and occupy the diagonal of the matrix. The input couplings M_{Sk} occupy the first row and column of the matrix from 1 to N. Similarly, the output couplings M_{Lk} occupy the last row and column of M. All other entries are zero (Iman, 2013; Saman,2013). Then, the elements of the transversal network are put in the following matrix as shown in Table 1

Table1: Transversal Matrix model

	S	1	2	3	⋯	K	⋯	N	L
S		T_{11}	T_{12}	T_{13}	⋯	T_{1K}	⋯	T_{1N}	K_O
1	T_{11}								T_{N1}
2	T_{12}								T_{N2}
3	T_{13}								T_{N3}
⋮	⋮								⋮
K	T_{1K}								T_{NK}
N	T_{1N}								T_{N1N}
L	K_O	T_{N1}	T_{N2}	T_{N3}	⋯	T_{NK}	⋯	T_{N1N}	

The coupling matrix obtained cannot be implemented practically. To obtain practical filters, a series of matrix rotations or similarity transformations will be applied to annihilate or make zero certain elements in the coupling matrix. The new matrix obtained should have the same electrical properties as the original.

The similarity transformations of an$N \times N$ coupling matrix [M] is mathematically defined as;

$$M_1 = R_1M_0R_1^1 \qquad (4)$$

Where [M0] is the original matrix and [M1] is the matrix after the operation. The rotation matrix [R] is defined by:

Table 2 : Similarity Transformation

	1	2	i		j		N
1	1						
2		1					
i			C_r		$-S_r$		
				1			
J			S_r		C_r		
						1	
N							1

$C_r = \cos\theta_r$, $Sr = \sin\theta r$ (5)

The rotation matrix [R] is an identity matrix, except that the elements $R_{ii} = R_{jj} = \cos\theta r$ and $R_{ji} = -R_{ij} = \sin\theta_r$, where [i,j] is the pivot of this rotation and θ_r is the angle of this rotation.

Microwave Resonator Design

(A) Assuming a Microwave filter is to be design at 2.327GHz centre frequency with f_1 and f_2 at 2.28GHz and 2.376GHz. Traansmission zeros are chosen at 2.2414GHz, 2.3015GHz, 2.3942GHz and 2.4314GHz. The fractional bandwidth of the microwave filter is 4.13%. Transmission zeros (P_s) in radian are –j1.8160,-j0.5351, j1.3801 and j2.1259.

The F_s is obtained using the Recursive method to obtain the polynomial

$F_s = S^8 + 2.4702S^6 + 1.9833S^4 + 0.5431S^2 + 0.02631.$

Roots of F_s also called the reflection coefficients are:

0.000000000000000 + j1.031960820284844
0.000000000000000 - j1.031960820284844
0.000000000000000 + j0.942817112810025
0.000000000000000 - j0.942817112810025
-0.000000000000000 + j0.674766844495440
-0.000000000000000 - j0.674766844495440
-0.000000000000000 + j0.247067729093188
-0.000000000000000 - j0.247067729093188

$E_s$is obtained using the equation $E_s . E_s^* = \frac{P_s^I P_s^I}{\epsilon_1^2} + F_s^I F_s^I$

The P_s, F_s and E_s are shown in Table 2 below:

Table 3: Polynomial functions for Microwave Resonator A

Coefficients	$\epsilon_1 = 6.1033$		
C_k	P_s	F_s	E_s
0	2.85103j	0.02631	0.49057
1	0	0	1.735
2	4.3372j	0.5431	2.9995
3	0	0	3.2876
4	j	1.9833	2.1536
5		0	1
6		2.4702	
7		0	
8		1	

(B) Assuming a 2649MHz Microwave resonator is to be designed, transmission zeros are obtained from the frequency chosen. Frequencies of 2540MHZ, 2680MHz, 2710MHz and 2720MHz are chosen. f_1 and f_2 are chosen at 2600MHz and 2700MHz. Transmission zeros in radian for 2540MHz, 2680MHz, 2710MHz and 2720MHz are obtained as –j2.2387, j0.6074, j1.1963 and j1.4960.

$P_S = js^4 + j3.9620s^2 - j2.4336$

Using the recursive method, F_S is obtained

$$F_S = s^8 + 0.50580s^6 + 0.9904s^4 + 0.5886s^2 - 0.03827$$

$E_S$is obtained using the equation (6) below

$$E_S.E_S^* = \frac{P_S^I P_S^I}{\epsilon_1^2} + F_S^I F_S^I \qquad (6)$$

To determine its transversal matrix for the microwave resonator filter at 2649MHz the following calculations below are carried out:

Transversal Matrix Calculations Using Microwave Resonator Filter B

Assuming ϵ_r=1 and return loss of 25dB, $\in_2 = 3.5812$ is obtained.

$$\varepsilon = \frac{1}{\sqrt{10^{RL/10} - 1}}.\ \frac{|P_{(s)}|}{|F_{(s)}|}|s = j$$

Where R_L is the prescribed Return loss

$$F_{(s)}.F_S^* + \frac{1}{\epsilon^2}P_{(s)}.P_{(s)}^* = E_{(s)}.E_{(s)}^*$$

$E_{(s)}.E_{(s)}^*$ =

-0.877860729454424 + 0.781688408311830i

-0.877860729454424 - 0.781688408311830i

-0.499045414018599 + 0.959668213703754i

-0.499045414018599 - 0.959668213703754i

0.000000000000000 + 1.073092626836047i

0.000000000000000 - 1.073092626836047i

0.877860729454424 + 0.781688408311831i

0.877860729454424 - 0.781688408311831i

0.499045414018598 + 0.959668213703754i

0.499045414018598 - 0.959668213703754i

-0.742776014979936 + 0.364287192497716i

-0.742776014979936 - 0.364287192497716i

-0.571431370480350 + 0.000000000000000i

0.742776014979935 + 0.364287192497714i

0.742776014979935 - 0.364287192497714i

0.571431370480350 + 0.000000000000000i

Using Hurwitz criterion for stability,

$$E_{(s)} = s^7 + 4.8106s^6 + 11.5015s^5 + 16.8995s^4 + 16.3544s^3 + 10.2722s^2 + 3.8212s^1 + 0.63219$$

Using the following equations,

m_1=Re$(e_0 + f_0)$ $+jim(e_1 + f_1)s^1 + Re(e_2 + f_2)s^2$ … … …

$n_1 = jim(e_0 + f_0)$ +Re$(e_1 + f_1)s^1$ +jim$(e_2 + f_2)s^2$ … … …

$y_{22} = \frac{m_1}{n_1}$ or $\frac{n_1}{m_1}$ depending on the order of the resonator

$y_{21} = \frac{\frac{P_S}{\epsilon}}{n_1}$ or $\frac{\frac{P_S}{\epsilon}}{m_1}$ depending on the order of the resonator

r_{22k}, r_{21k} and λ_k are obtained.

$r_{22k} =$

$7.916899486401720 - 0.000000000000001i$

$7.916899486401720 + 0.000000000000001i$

$0.285985092537727 - 0.000000000000000i$

$0.285985092537726 + 0.000000000000000i$

$0.191209561395817 + 0.000000000000000i$

$0.191209561395817 - 0.000000000000000i$

$0.168511719329477 + 0.000000000000000i$

$r_{21k} = 0.000000000000000 - 0.002755620139700i$

$-0.000000000000000 - 0.002755620139700i$

$0.000000000000000 - 0.011167077534656i$

$-0.000000000000000 - 0.011167077534656i$

$-0.000000000000000 + 0.029672864563542i$

$0.000000000000000 + 0.029672864563542i$

$0.000000000000000 - 0.031500333778371i$

λ_k=0.000000000000000 + 6.941273311711340i

0.000000000000000 - 6.941273311711340i

0.000000000000000 + 1.199423483108557i

0.000000000000000 - 1.199423483108557i

0.000000000000000 + 0.557860305355764i

0.000000000000000 - 0.557860305355764i

0.000000000000000 + 0.000000000000000i

The P_s, F_s and E_s are shown in Table 4 below:

Table 4: Polynomial functions for Microwave Resonator B

Coefficients	$\in_2 = 6.5977$		
C_k	P_s	F_s	E_s
0	-2.4336j	-0.03827	0.63219
1	0	0	3.8212
2	3.9620j	0.5886	10.2722
3	0	0	16.3544
4	J	0.9904	16.8995
5		0	11.5015
6		0.5058	4.8106
7		0	1
8		1	0

(C) The same is done for microwave filter design at 2.42GHz with transmission zeros chosen at are –j1.5322, j0.5406, j1.3389 and j2.3613 and $\in_3 = 3.5812$ The P_s, F_s and E_s are shown in table 4 below

The three designed filters are summarized in the table below:

.Table 5: Polynomial functions of a Novel method of polynomial functions design for triple band resonator filter.

C_k	P_s^I	P_s^{II}	P_s^{III}	F_s^I	F_s^{II}	F_s^{III}	E_s
0	2.8510j	-2.6187	- 2.4336j	0.02631	- 0.02236	- 0.03827	0.4905
1	0	0	0	0	0	0	2.7545
2	4.3372j	1.3358j	3.9620j	0.5431	0.22435	0.5886	7.2363
3	0	0	0	0	0	0	11.5718
4	j	J	j	1.9833	0.07161	0.99041	12.2423
5				0	0	0	8.7932
6				2.4702	1.2483	0.50581	3.9909
7				0	0	0	1
8				1	1	1	0

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C_k	$P_s = \frac{P_s^I}{\in_1} + \frac{P_s^{II}}{\in_2} + \frac{P_s^{III}}{\in_4}$	$F_s = F_s^I + F_s^{II} + F_s^{III}$	E_s^*
0	-0.6094j	-0.0343	0.4905
1	0	0	2.7545
2	2.0194j	1.3561	7.2363
3	0	0	11.5718
4	-0.5946	3.0453	12.2423
5		0	8.7932
6		4.2243	3.9909
7		0	1
8		3	0

Note:

$$E_s.E_s^* = \frac{P_s^I P_s^I}{\epsilon_1^2} + F_s^I F_s^I + \frac{P_s^{II} P_s^{II}}{\epsilon_2^2} + F_s^{II} F_s^{II} + + \frac{P_s^{III} P_s^{III}}{\epsilon_3^2} + F_s^{III} F_s^{III} (7)$$

The same method used to determine the transversal matrix of microwave resonator filter B is applied to derive the transversal matrix of the three combined microwave resonator filters.

The transversal matrix of the three combined microwave resonator filters after undergoing optimization using the swarm particle optimization cost function of equation 8 yields the transversal matrix M.

$$\sum_{i=1}^{N} |S_{11}(w_{zi})|^2 + \sum_{i=1}^{P} \left|S_{21}(w_{pi})\right|^2 + \sum_{i=1}^{Q} \frac{1}{\varepsilon} \left|\frac{S_{11}(w_{zi})}{S_{21}(w_{pi})}\right|^2 + \sqrt{\frac{1}{4M} \sum_{i=1}^{M} \left|\frac{S_{21}(\omega_{pi})}{H(\omega_{pi})}\right|^2 + \frac{1}{4N} \sum_{i=1}^{N} \left|\frac{S_{11}(\omega_{zi})}{1-|H(\omega=p_i)|^2}\right|^2} \quad (8)$$

M=

$$\begin{matrix}
0 & 0.0023 & 0.0023 & 0.04157 & 0.04157 & 0.1088 & 0.1088 & 0.1121 & 0.348 \\
0.0023 & 7.1817 & 0 & 0 & 0 & 0 & 0 & 0 & 3.3608 \\
0.0023 & 0 & -7.1817 & 0 & 0 & 0 & 0 & 0 & 3.3608 \\
0.04157 & 0 & 0 & 1.0646 & 0 & 0 & 0 & 0 & 0.4476 \\
0.04157 & 0 & 0 & 0 & -1.0646 & 0 & 0 & 0 & 0.4476 \\
0.1088 & 0 & 0 & 0 & 0 & 0.4972 & 0 & 0 & 0.4357 \\
0.1088 & 0 & 0 & 0 & 0 & 0 & -0.4972 & 0 & 0.4357 \\
0.1121 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.3762 \\
0.348 & 3.3608 & 3.3608 & 0.4476 & 0.4476 & 0.4357 & 0.4357 & 0.3762 & 0
\end{matrix}$$

The folded matrix obtained after novel similarity transformation was carried out is:

0	1.14303	0	0	0	0	0	0	2.3925
1.14303	8.5658	4.5947	0	0	0	0	0	0
0	0.69634	-1.1651	0.28007	0	1.1934	0	-0.08369	0
0	0	-0.5565	0.122	-0.54189	0	0	0	0
0	0	0	-0.2336	0.06288	0	0	0	0
0	0	1.1934	0	0	0.2745	0.7429	0	0
0	0	0	0	0	0.7429	-0.02035	0	0
0	0	-0.08369	0	0	0	0	-0.67114	0
2.3925	0	0	0	0	0	0	0	0

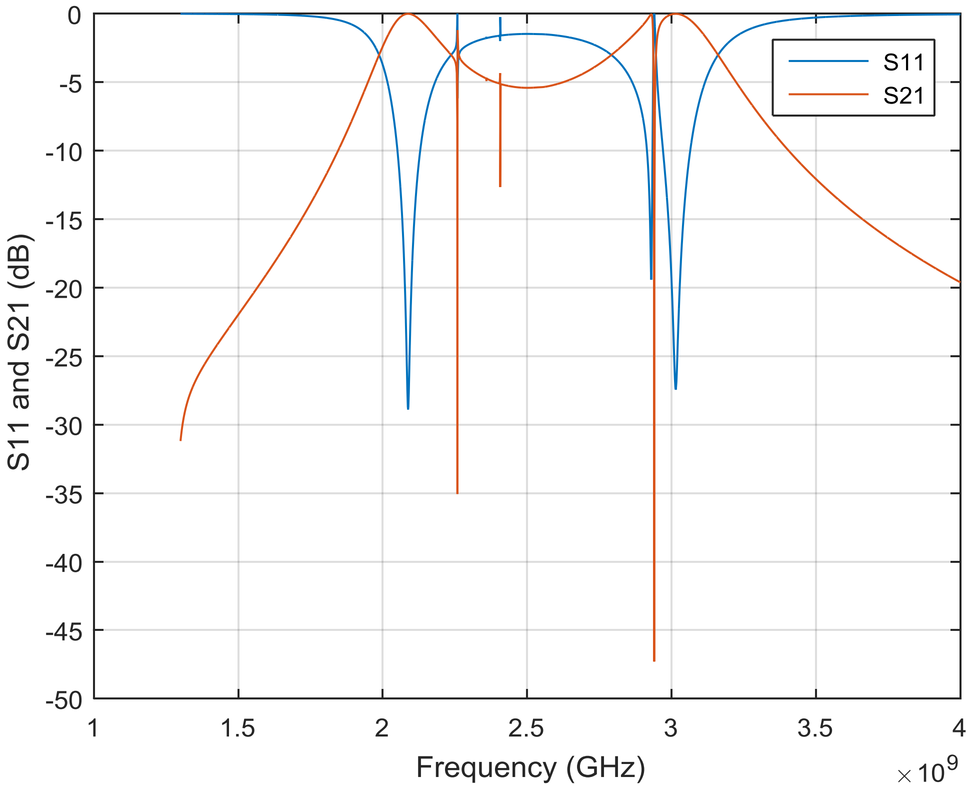

Figure 2: Transmission and reflection coefficients ($S_{21}$and S_{11}) of the folded matrix

Theory of Physical Microwave Filter Design

If a half wavelength micro strip line of 50Ω impedance is designed at 2.32GHz ,2.42 and 2.55GHz designed on a RT/Duroid substrate having a thickness h = 1.27mm with relative dielectric constant ε_{re} = 6.15. Loss tangent (tan δ) is 0.0027.

The width of a 50Ω micro strip line is estimated to be

W= 1.98mm

From equation (i), the effective electric constant (ε_{re}) is

$$\varepsilon_{re} = 4.4481$$

The resonant line can be calculated as

$$l = \frac{\lambda g}{2} = \frac{V_p}{2f} = \frac{c}{2f\sqrt{\varepsilon_{re}}} \quad (9)$$

$$\beta = \frac{2\pi f\sqrt{\varepsilon_{re}}}{c} \quad (10)$$

Physical Design

The thickness is 0.0745mm with a dielectric constant of 6.15 and loss tangent of 0.0002. The design is implemented using microstrip technology

The square open loop resonator physical dimension was designed on frequencies of 2.48GHz, 2.92GHz and 3.3GHz.

The external coupling coefficient is given by

$$R = \frac{\Delta f \pm 90°}{\Delta f} \quad (11)$$

The spacing between two resonators is determined by inter-resonator coupling coefficients $M_{ij} = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2}\frac{f_0}{\Delta f}$ (12)

Where f_1 and f_2 are lower and higher frequencies at which the peak is observed.

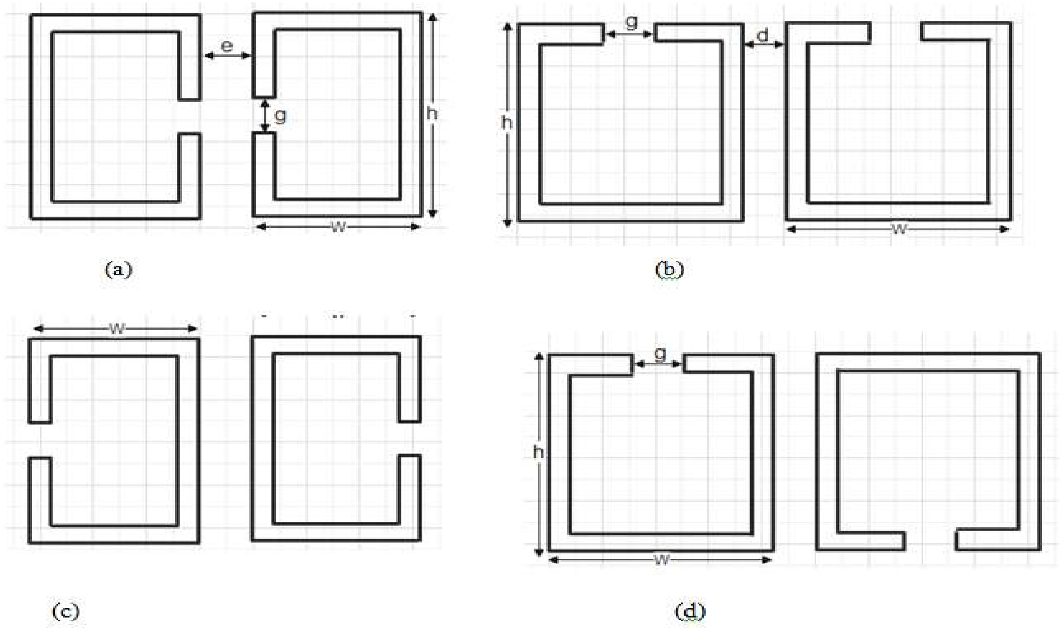

Figure 3: Coupling structure of square open loop resonator (a) electric coupling structure (b) mixed-coupling structure (c) magnetic coupling structure (d) mixed-coupling structure

Novel Physical Design

A novel square open loop resonator filter was designed that is able to remove spurious harmonics corresponding with the frequency of the inner resonator structure. The novel resonator filter inner resonator was designed to operate on 2.37GHz based on its length of 29.99mm with the outer resonator frequency to operate on 1.78GHz based on its length of 39.99mm.

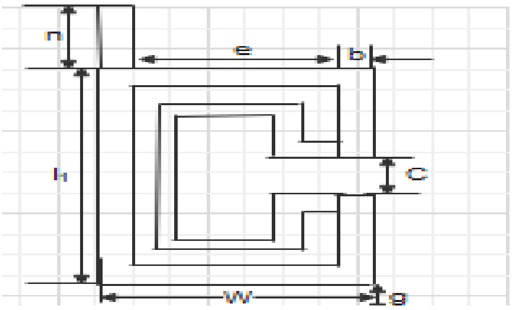

Figure 4: one resonator Square open loop resonator filter

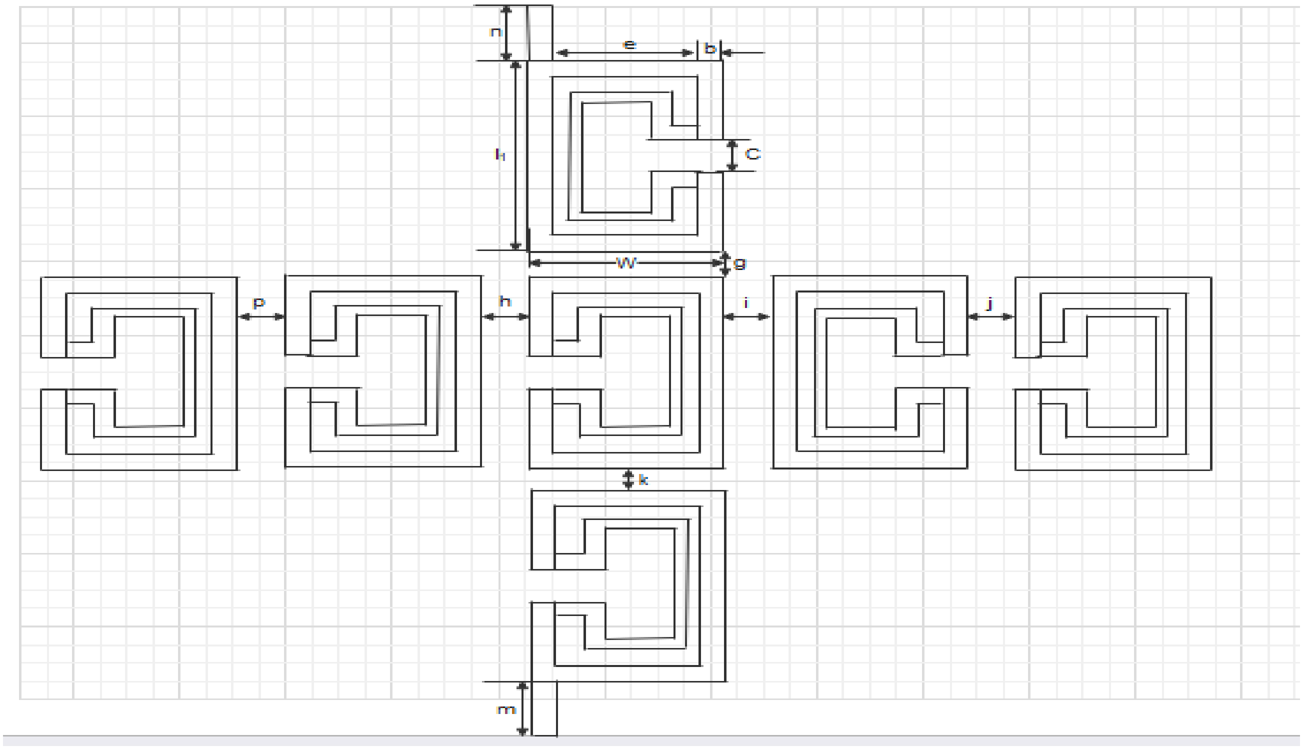

Figure 5: layout structure of a seven resonator square open loop resonator filter

High Frequency Structure Simulator Software

High frequency structure simulator software was used to simulate the bandpass resonator filter. Various shape dimension of the design is drawn with ports added on either side. Lumped ports are used if the port is placed in the substrate and wave ports are employed if the ports used are not placed inside the substrate, both rather at its sides. Then validation check is performed, followed by analyze all to do the necessary simulation.

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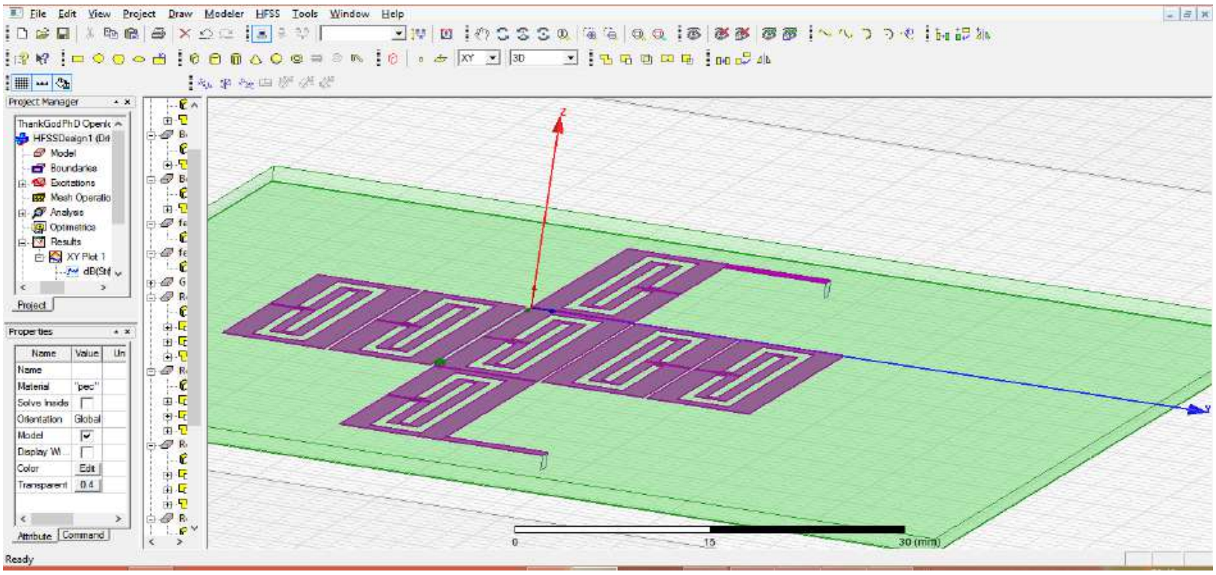
RESULTS AND DISCUSSION

Table 6: Physical dimensions of the square open loop resonator filter (outer square)

Parameters	Values (mm)
b	1.98
C	0.3961
e	4.1
g	0.4213
i	0.3641
j	0.3341
k	0.3025
li	39.99
t	0.0745
m	8.06
n	8.06

Table 7: Physical dimensions of the square open loop resonator filter (inner square)

Parameters	Values(mm)
b	0.99
Inner e	2.78
C	0.3961
li	29.99
t	0.0745
m	8.06



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Name		
Material	"pec"	
Solve Inside	☐	
Orientation	Global	
Model	☑	
Display W...	☐	
Color	Edit	
Transparent	0.4	

Figure 6: physical structure of the novel square open loop resonator filter

Figure 6 above shows the physical structure of the square open loop resonator to be simulated using High frequency simulation software. After validation check is performed, showing that the design settings, 3 dimensional model, boundaries and excitation, frequency analysis set up, optimetrics and radiation are in order, simulation of the entire structure starts. After validation, the result of the simulation is shown in figures 7,8,9, 10, 11,12.

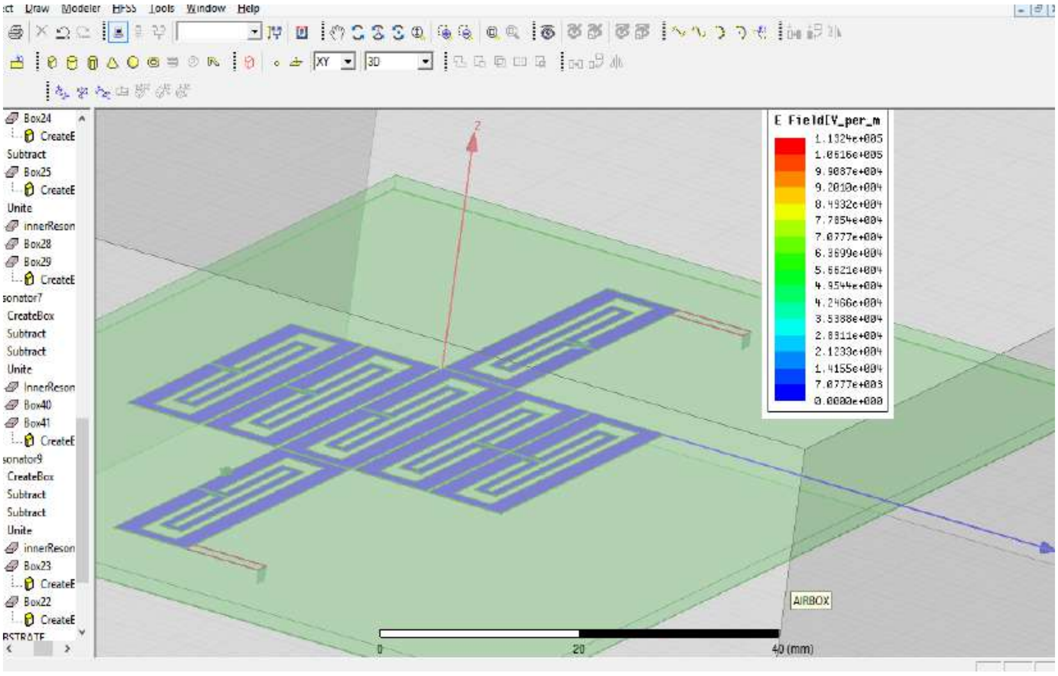


Figure 7: Electric field pattern of the simulated microwave resonator filter

The electric field per metre bar value shows different colouration with the dark orange having the highest value of 1.1324 $\times e^5\ v/m$ and the lowest volt per metre value is the dark blue colouration having a value of 7.0777 $\times e^3\ v/m$. The electric field of the square open loop resonator is of the dark to light blue colouration which is in order.

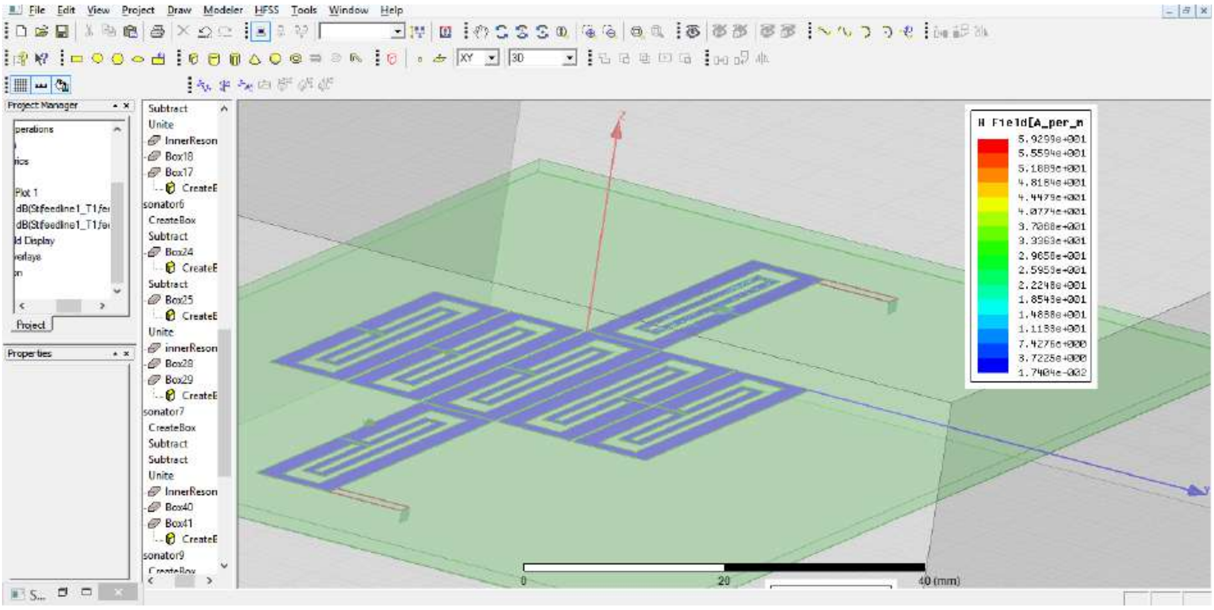


Figure 8: Magnetic field pattern of the simulated microwave resonator filter

The magnetic field pattern of figure 8 shows that each colouration corresponds to different values measured in ampere/metre. The dark orange has the highest value of $4.5005 \times e^{1}\, A/m$ and the least value being the dark to light blue colouration ranging from $1.740 \times e^{-2}\, A/m$ to $2.8292 \times e^{0}\, A/m$. The magnetic field pattern colouration of the microstrip resonator filter shows that the dark to light blue colouration dominate.

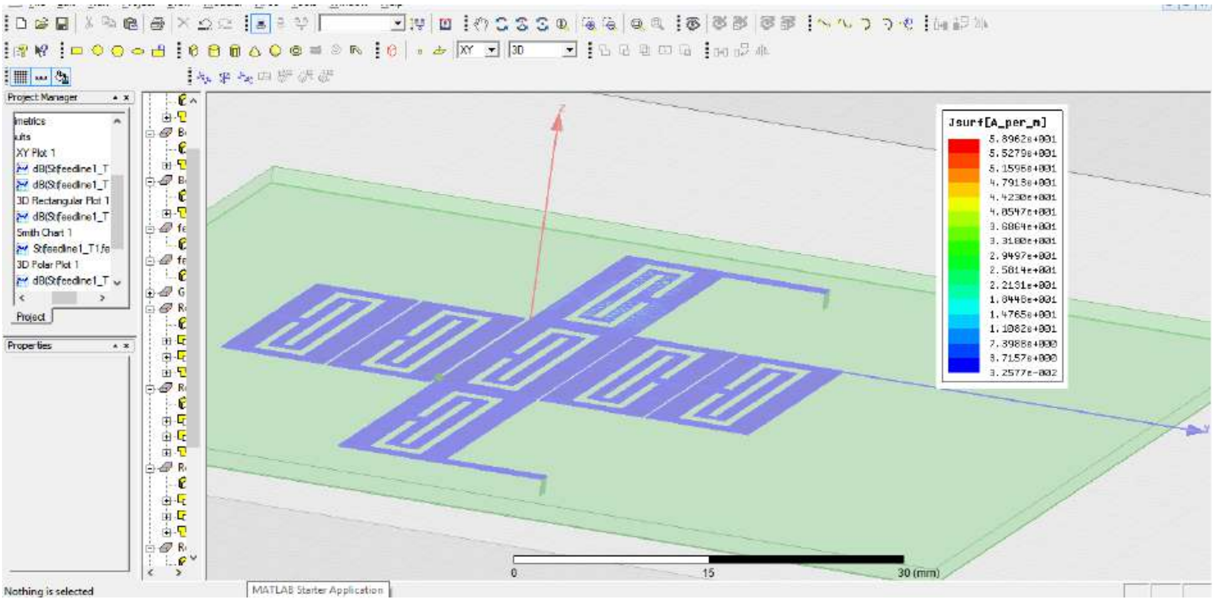
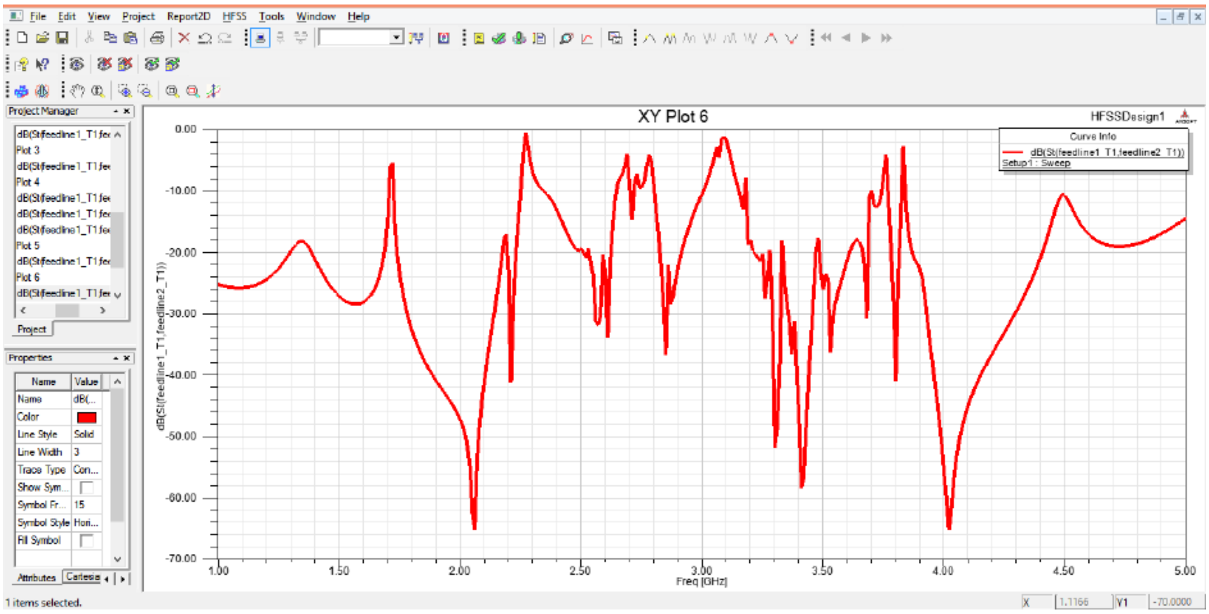


Figure 9: current density of the simulated microwave resonator filter

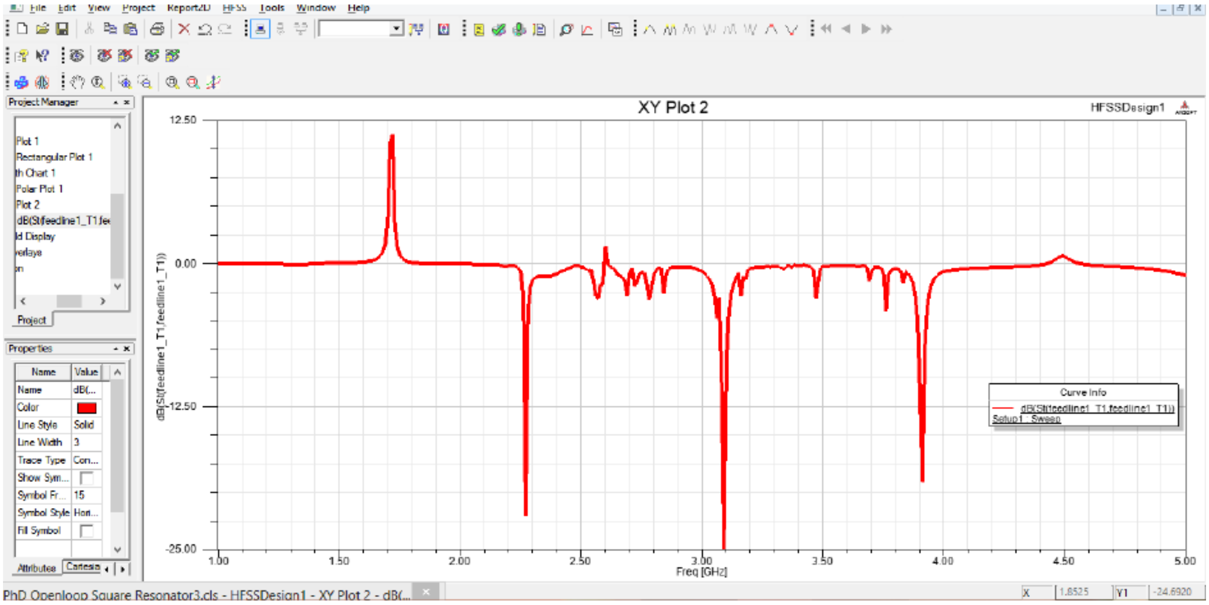
The surface current density of the square open loop resonator filter shows different shades of colouration corresponding to different values in A/m. The highest value being the dark orange colouration corresponding to a value of $6.9663 \times e^{1}\, A/m$ while the dark blue colouration has the least value of $3.2577 \times e^{-2}\, A/m$. The surface current density of the square open resonator filter shows basically the light to dark blue colourations and the values are accurate or in order for this resonator filter.



Name	Value
Name	dB(...
Color	🟥
Line Style	Solid
Line Width	3
Trace Type	Con...
Show Sym...	☐
Symbol Fr...	15
Symbol Style	Hori...
Fill Symbol	☐

Figure 10: The transmission coefficient S_{21} of the microwave resonator filter.

The transmission coefficient shows passbands of 2.23GHz and 3.2GHz with passband of 2.7GHz attenuated or pushed down. This value is consistent with the value of the folded matrix in figure 2 used to physically implement the designed resonator filter having passbands of 2.25GHz and 3.2GHz with 2.65GHz pushed down.



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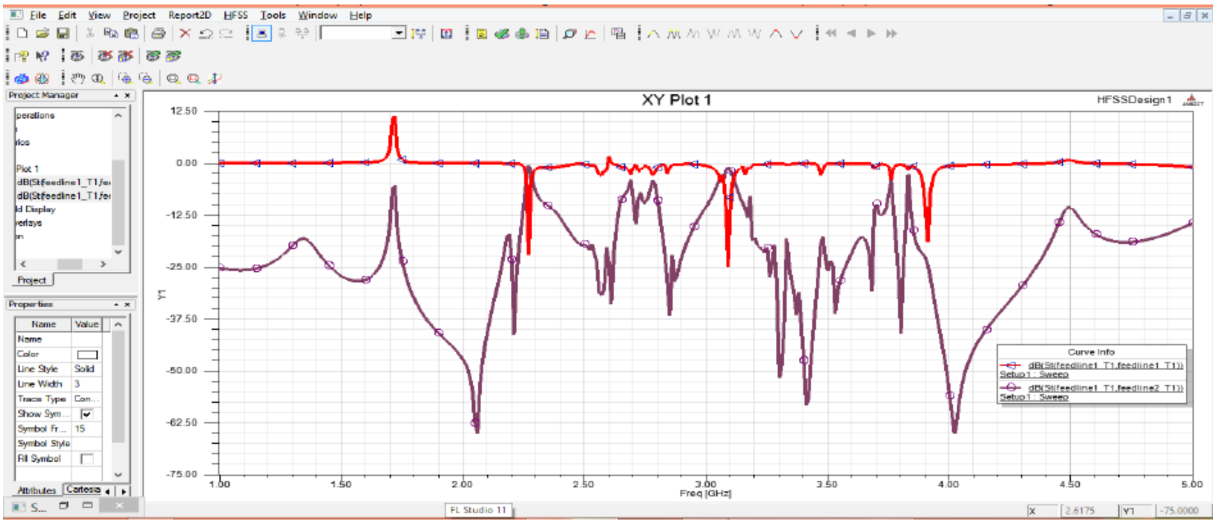
PhD Openloop Square Resonator3.cls - HFSSDesign1 - XY Plot 2 - dB(...

Figure 11: The reflection coefficient S_{11} of the microwave resonator filter

The above figure 11 shows the reflection coefficient of close to 20dB for the different pass bands of the microwave resonator filter.



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Name	
Color	☐
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Figure 12: wideband view of the transmission coefficient and reflection coefficient of the novel square open loop resonator filter showing all harmonics has been successfully removed.

Figure 12 shows the transmission and reflection coefficient being represented by the red and brown colour of the frequency response. It shows that the transmission coefficient indicated by the brown colouration and the reflection coefficient being represented by the red colouration. Careful study of the design shows that the outer resonator has a length of 39.99mm which is equivalent to frequency of 1.782GHz using microstrip technology. Here, frequency corresponding to the outer resonator is being removed or eliminated. The inner resonator length which is the frequency of interest has a length of 29.99mm corresponding to a frequency of 2.3GHz using microstrip technology. Note that a tripleband on figure 2 was physically implemented using microstrip technology with first passband at 2.2GHZ, second pass band not properly formed and the third pass band at about 3.2GHz.

It is also observed that all other frequencies (or harmonics) have been successfully removed. Changing the frequency ranges from the one used in Figure 12 ranging from 1GHz to 5GHz to one ranginging from IGHz to value higher than 5GHz shift the entire passband frequency to another passband frequency indicating that at the initial pass band frequency, no other frequency are present within the passband range and therefore all harmonics was successfully removed.

DISCUSSIONS

The above figure 12 shows that all harmonics has been successfully removed. Changing the frequency range more than 5GHz will cause a shift of the centre frequency moving from to another centre frequency showing all harmonics has been successfully removed.

CONCLUSION

Spurious harmonics and unwanted frequencies poses great problem to effective mobile communication and are therefore acts as threat to good signal reception. The presence of this unwanted frequencies and harmonic usually causes interferences in mobile communication system. Unwanted frequencies that will interfere with our choice frequency or frequencies of communication are being successfully removed using these novel design of the square open loop resonator employing both outer and inner resonators.

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