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Research Article

Microstrip Tunable Bandpass Filter with the Colinear Resonators

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This paper presents the lumped element circuit and transmission line equivalent circuit for a varactor-tuned bandpass filter. The filter consists of transmission lines, fixed capacitors, and a varactor diode. The colinear resonant sections, in this filter, are not configured in parallel, as they are in a conventional combline filter. For this reason the overall area of the filter is reduced. The passband of the filter can be tuned from 0.69 GHz to 1.20 GHz by varying the capacitance of the varactor diode. The insertion loss of this filter changes from 1.2 dB to 2.1 dB across this bandwidth.

1. Introduction

Tunable bandpass filters will play an important role in the future communication [1–3]. Compact, highly selective, and tunable filters with low insertion loss are widely required. Different tuning technologies have been developed, including varactor diode, mechanical tuning, microelectromechanical (MEMS) component, and p-i-n diode [4–10]. One of the most popular techniques for achieving continuous tuning by electrical means within a bandpass filter is to use a varactor diode. Hunter and Rhodes [11] report a tunable combline bandpass filter which can be tuned from 3.2 GHz to 4.9 GHz with a 3–5 dB insertion loss. Sánchez-Renedo et al. [12] present a developed combline filter structure with a continuous tunability for both the center frequency and bandwidth. El-Tanani and Rebeiz [13] describe a combline filter based on corrugated microstrip coupled line which offers miniaturized size. This filter has a frequency coverage of 1.32–1.89 GHz with constant passband bandwidth across the tuning range. Tang and Hong [14] present a novel tunable bandpass filter based on a dual mode microstrip open-loop resonator. This filter offers constant absolute bandwidth and wide tuning range.

Most of the designs reported to date are based around coupled lines. Primarily these structures are coupled together magnetically. The radiation loss of the coupled line is high when the coupling is weak. In some application, it is not easy to establish the coupling matrix. This is the case for coplanar waveguide (CPW), for example.

This paper presents a novel tunable bandpass filter with colinear resonators. The filter is not based around parallel coupled lines. The overall area of this filter is smaller than the conventional combline filter. Equivalent circuits of the new filter are analyzed.

2. Equivalent Circuit

Figure 1 shows the proposed filter. The filter consists of microstrip lines, fixed capacitors, and a varactor diode. The main resonator is composed of a varactor diode and microstrip line L_2.

A lumped element equivalent circuit has been derived for this filter. The equivalent circuit is shown in Figure 2. The basic resonator is represented by L_{i2} and capacitance of the varactor (C_2). The external coupling is controlled by the "L" shaped circuit located on either side of this resonator.

The circuit element values are presented in Table 1. The S parameters of the lumped element equivalent circuit are calculated in MATLAB 6.5, as shown in Figure 3. It can be observed from Figure 3 that the centre frequency of the passband can be tuned from 0.58 GHz to 1.26 GHz by varying

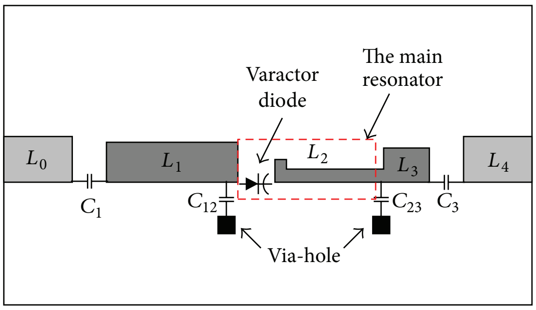


Figure 1: The tunable bandpass filter.

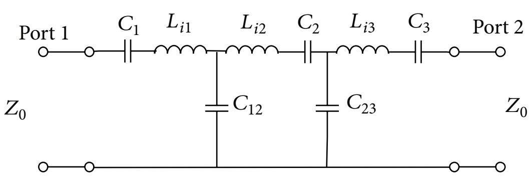


Figure 2: Lumped element equivalent circuit.

Table 1: Equivalent circuit element values.

Element	Value
C_1	3.3 pf
C_2	C_v
C_3	3.3 pf
C_{12}	6.8 pf
C_{23}	6.8 pf
L_{i1}	5 nH
L_{i2}	20 nH
L_{i3}	5 nH

the capacitance of C_2 from 15 pf to 1 pf. The absolute 3 dB bandwidth of this filter is 100 ± 2 MHz across the tuning range. According to computer simulation the return loss is better than 20 dB across the entire tuning range.

Figure 4 shows a transmission line equivalent circuit developed from the lumped element equivalent circuit. The capacitors retain their value and position in the transmission line circuit. The inductors L_{i1}, L_{i2}, and L_{i3} in the lumped element circuit are replaced by sections of transmission line TX_1, TX_2, and TX_3, respectively. Transmission through the filter, that is, S_{21}, can be deduced with reference to the *ABCD* matrix.

The *ABCD* matrix for the series capacitor C_1 is

$$A_1 = \begin{bmatrix} 1 & Z_{\text{cap1}} \\ 0 & 1 \end{bmatrix}, \tag{1}$$

where $Z_{\text{cap1}} = 1/j\omega C_1$.

The *ABCD* matrix of transmission line TX_1 is

$$A_2 = \begin{bmatrix} \cos\beta l_1 & jZ_1 \sin\beta l_1 \\ jY_1 \sin\beta l_1 & \cos\beta l_1 \end{bmatrix}, \tag{2}$$

where Z_1, Y_1 are the characteristic impedance and admittance of the transmission line, β is the transmission line phase constant, and l_1 is the length of the transmission line.

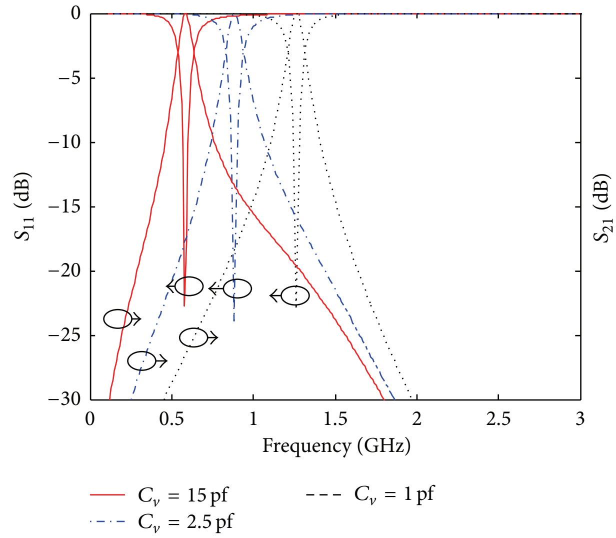


Figure 3: Calculated S parameters.

Figure 4: Transmission line equivalent circuit.

The *ABCD* matrix of the parallel capacitor C_{12} is

$$A_3 = \begin{bmatrix} 1 & 0 \\ Y_{\text{cap12}} & 1 \end{bmatrix}, \tag{3}$$

where $Y_{\text{cap12}} = j\omega C_{12}$.

The *ABCD* matrices of the other capacitors and transmission lines follow in the same way, respectively.

According to matrix algebra the *ABCD* matrix of the whole circuit shown in Figure 3 can be described as follows:

$$A = A_1 A_2 A_3 A_4 A_5 A_6 A_7 A_8 = \begin{bmatrix} a & b \\ c & d \end{bmatrix}. \tag{4}$$

The S parameter matrix can be derived from the *ABCD* matrix as follows:

$$S = \begin{bmatrix} \dfrac{a + b/Z_0 - cZ_0 - d}{a + b/Z_0 + cZ_0 + d} & \dfrac{2(ad - bc)}{a + b/Z_0 + cZ_0 + d} \\ \dfrac{2}{a + b/Z_0 + cZ_0 + d} & \dfrac{-a + b/Z_0 - cZ_0 + d}{a + b/Z_0 + cZ_0 + d} \end{bmatrix}. \tag{5}$$

The lengths of the transmission line sections TX_1, TX_2, and TX_3 were set to 0.12, 0.14, and 0.03 wavelengths at 1 GHz, respectively. Normally one would expect the length of the input and output lines of a filter to be identical. In this case, however, the lengths of TX_1 and TX_3 are different. This is mainly caused by the asymmetry of the main resonator (formed by TX_2 and C_2). The length of TX_3 is set shorter than that of TX_1. This makes the passband bandwidth constant, across the entire tuning range. The characteristic impedance of TX_1 and TX_3 is set to 50 Ω. The characteristic impedance of TX_2, however, is set to 100 Ω. This is necessary in order to achieve a high inductance value from a short length of transmission line.

Figure 5 shows the calculated results from (5) using MATLAB 6.5. It can be observed that the centre frequency of the passband can be tuned from 0.65 GHz to 1.24 GHz by varying the capacitance of C_2 from 15 pf to 1 pf. The absolute 3 dB bandwidth of this filter varies from 100 MHz to 97 MHz across the tuning range. The 3 dB bandwidth is nearly constant as the center frequency of the passband is changed.

3. Microstrip Line Filter Design

A microstrip line tunable bandpass filter is developed based on transmission line circuit shown in Figure 4. The structure of the filter is shown in Figure 1. L_0 and L_4 are 50 Ω transmission lines. Both of these lines are 3.38 mm wide and 10 mm long. However the length of L_0 and L_4 has no impact on the performance of this design. L_1, L_2, and L_3 are 3 resonant sections. The length and width of these 3 resonant sections are optimized to provide constant bandwidth across the whole tuning range. L_1 is 3.05 mm wide and 24.6 mm long. L_2 is 0.46 mm wide and 20 mm long. A 2 mm by 2 mm pad is introduced at the end of L_2 nearest L_1. This pad provides a surface on which to solder the varactor diode. L_3 is 2.44 mm wide and 8.2 mm long. The resonator within this filter is 52.8 mm long (i.e., nearly $\lambda_{\text{eff}}/4$ at 1.2 GHz). Two chip capacitors, labeled C_1 and C_3 (ATC 600S, C = 3.3 pf), are used to connect the transmission lines L_0 and L_4 to L_1 and L_3 together. A second pair of chip capacitors, labeled C_{12} and C_{23} (ATC 600S, C = 6.8 pf), are used to connect L_1 and L_2 to the grounding vias.

Two 10 kΩ RF-choke resistors are used to connect the resonator to the bias line. The filter is fabricated on a Taconic TLY-3 substrate (i.e., h = 1.14 mm, ε_r = 2.33, and $\tan\delta$ = 0.0009). An MV31009 surface mount varactor diode, from MDT Corporation, was used in the prototype circuit. The relationship between the voltage and the capacitance is given in Table 2.

The equivalent series resistance of the varactor is 0.455 Ω. The S parameters of the microstrip line filter can be calculated using (5), as mentioned earlier. The $ABCD$ matrix of the varactor diode can be represented as follows:

$$A_6 = \begin{bmatrix} 1 & \dfrac{1}{j\omega C_2} + 0.455 \\ 0 & 1 \end{bmatrix}. \tag{6}$$

Figures 6(a) and 6(b) show the scattering parameters obtained through simulation using CST MWS. The filter's passband centre frequency can be tuned from 0.69 GHz to 1.24 GHz by varying the reverse voltage applied to the varactor diode (C_2). The absolute 3 dB bandwidth of this filter varies from 125 MHz to 136 MHz as the filter is tuned. The insertion loss is nearly 0.8 dB, and the return loss is better than 18 dB across the whole tuning range.

The capacitances of C_{12} and C_{23} control the loaded Q-factor of the filter (see Figure 1). Increasing the capacitance increases the loaded Q-factor. In order to minimize the total number of varactors required and hence the overall insertion loss we demonstrate a 1-pole tunable bandpass filter. In a multipole bandpass filter these capacitors would provide a convenient way by which to vary the coupling coefficient over a wider range of values. It is also possible to control the bandwidth of the passband by tuning the coupling capacitance.

Table 2: Capacitance versus reverse voltage (MV31009).

Voltage (V)	Capacitance (pf)
0	22.08
3	4.77
5	3.01
7	2.18
9	1.71

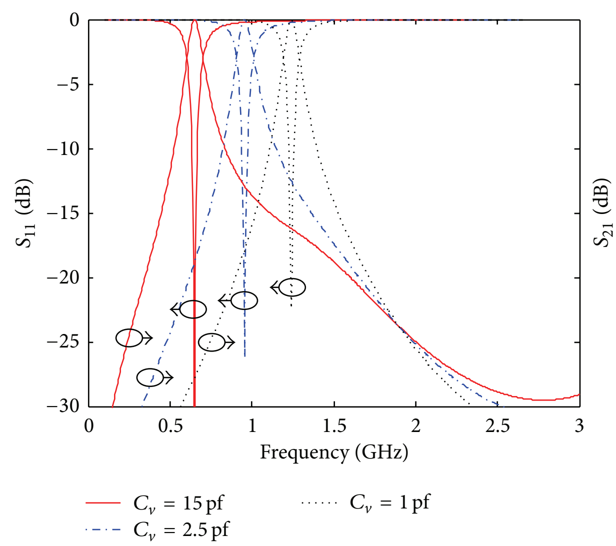
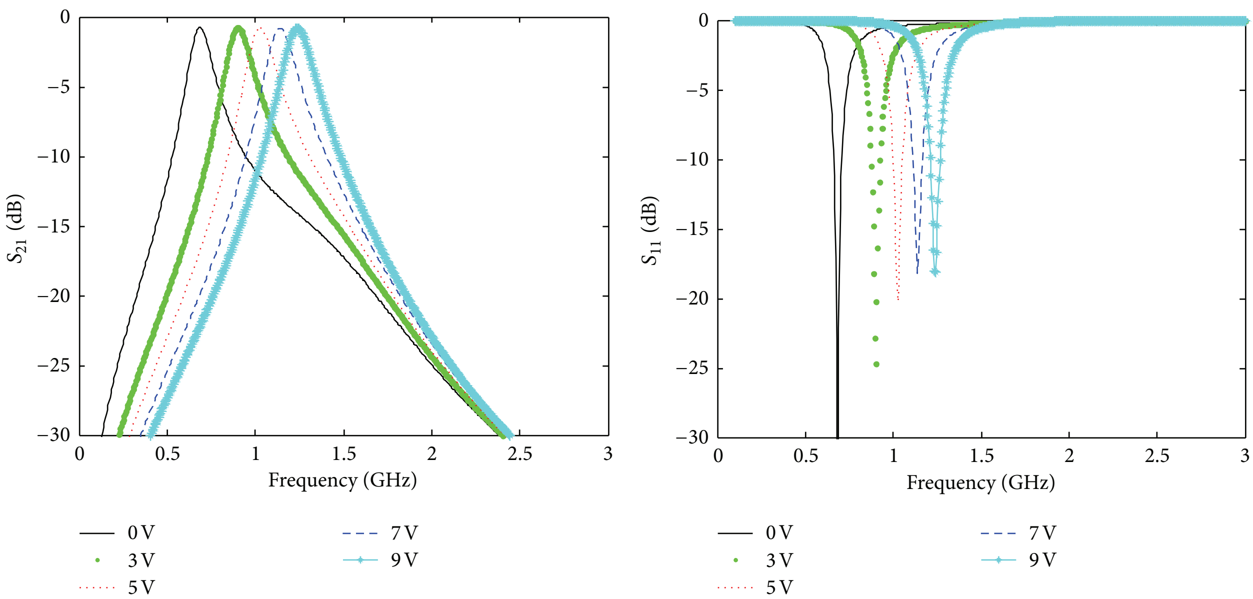


Figure 5: Calculated S parameters.

4. Measurement Results

Figure 7 shows a prototype of the tunable bandpass filter. Two SMA connectors are soldered on each side of the filter. The reverse voltage is applied across the bias lines depicted in Figure 7. Agilent 8720 VNA is used in this measurement. Figure 8(a) shows measured S_{21} results. The centre frequency of the passband can be tuned from 0.69 GHz to 1.20 GHz. The insertion loss varies from 1.2 dB to 2.1 dB as the filter is tuned throughout this range. The insertion loss is suppressed below −30 dB from 0.1 GHz to 2.5 GHz. According to the measurement results the 3 dB bandwidth of the filter changes from 118 MHz to 69 MHz, as the filter is tuned throughout its full range. Compared with the simulation results, the measurement results indicate that the passband bandwidth of the filter becomes narrower as the filter is tuned towards higher frequencies. The insertion loss observed through the measurement is also higher than that predicted by simulation. These differences are mainly due to an increase in the parasitic resistance and inductance of the fixed capacitors and the varactor diode. The simulation results show that the tuning range is restricted by the variation in the capacitance of the varactor diodes.



(a) (b)

Figure 6: Simulated S parameters.

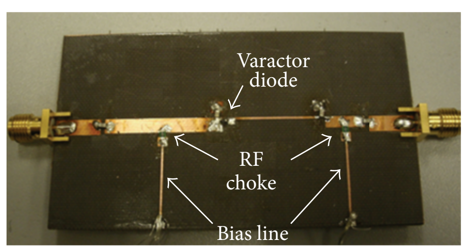
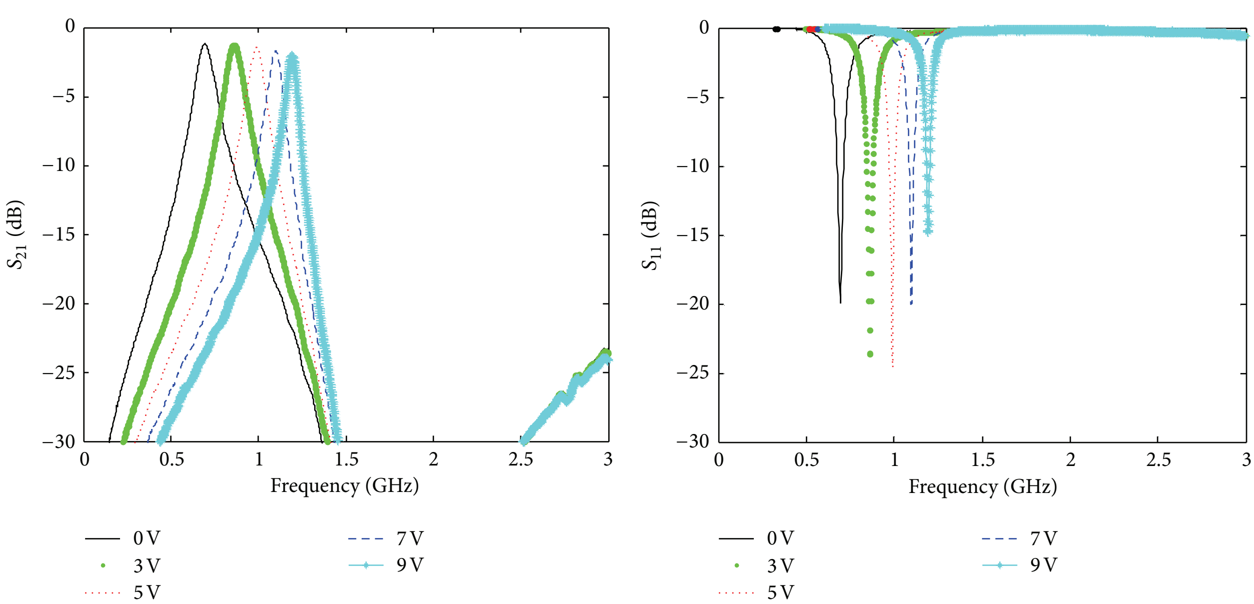


Figure 7: Prototype of the filter.



(a) (b)

Figure 8: Measured S parameters.

Figure 8(b) shows the measured return loss. Inspection of the results reveals that the return loss is better than 14.8 dB for all of the tuning states demonstrated.

5. Conclusion

This paper presents a new form of tunable bandpass filter with colinear resonators. Unlike many of the other tunable filters in the literature the filter is not based around coupled lines. The equivalent circuits of the new filter are analyzed. Experimental results show that the passband centre frequency can be tuned from 0.69 GHz to 1.20 GHz by varying the capacitance of the varactor from 22.08 pf to 1.71 pf.

Conflict of Interests

The authors declare that they have no conflict of interests.

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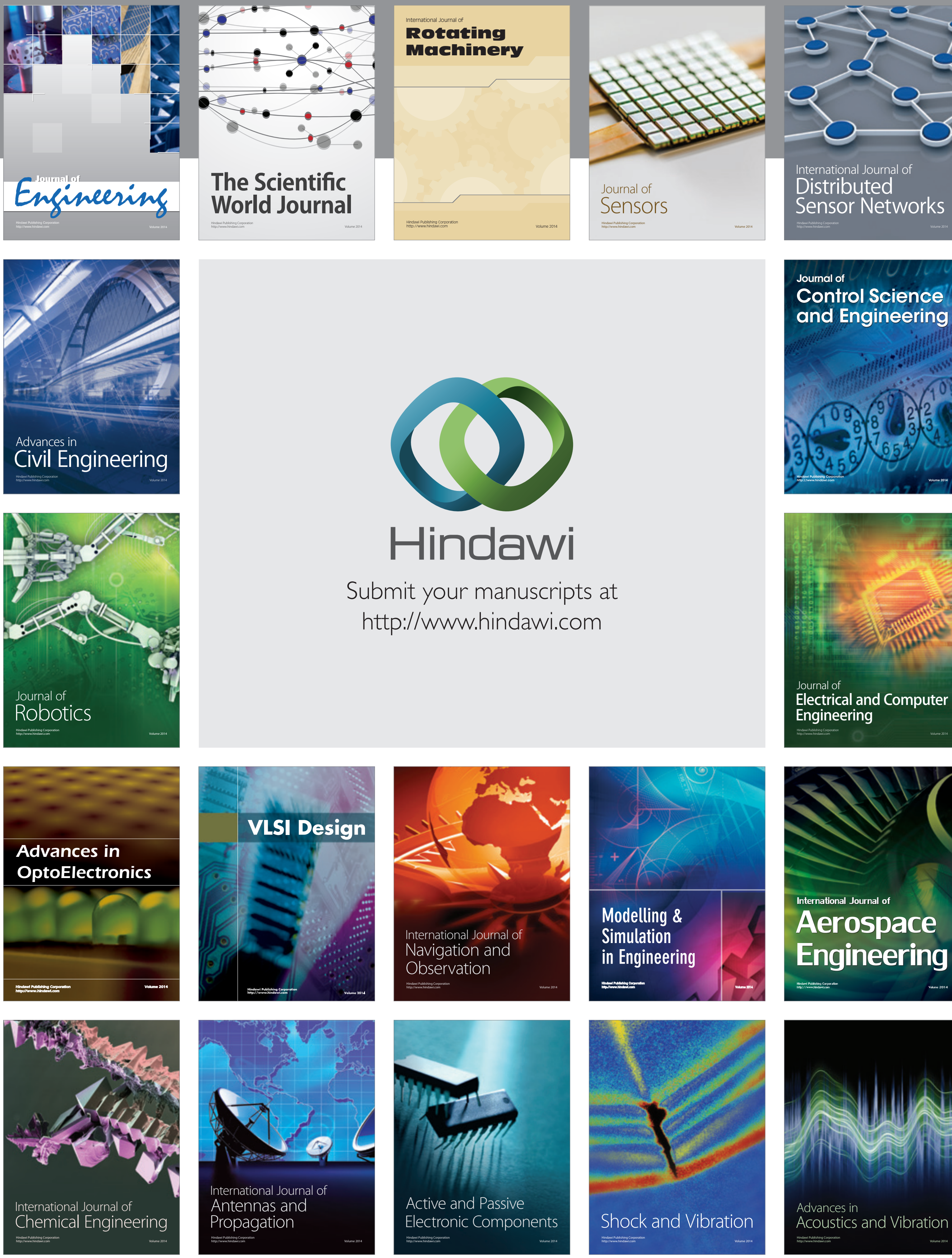
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