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PULSE INJECTION FOR REAL-TIME MONITORING OF TUNABLE HIGH-Q FILTERS

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found from Eq. (1) corresponds to the insertion phase divided by the length of the device, regardless of whether or not the device is dispersive and has significant attenuation. In addition, the attenuation coefficient $\alpha$ is accurately computed from the simplified form of Eq. (1) that appears in the second line of Eq. (4).

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ABSTRACT: A new measurement technique for monitoring the frequency of high-Q microwave tunable filters in real time is presented in this work. Relying on the natural response of a cavity resonator, the technique provides the operating frequency information in a binary digital format, making it easy to read and process. The presented methodology is experimentally demonstrated on hand-stop filters, implemented by highly loaded substrate-integrated cavities in the 1.5 GHz range. Measured results show that the changes in frequency as small as 2–20 MHz (0.13–1.3%) can be detected for a tunable cavity filter (1.42–1.65 GHz) within 1 ms, based on commercial-off-the-shelf components. © 2014 Wiley Periodicals, Inc. Microwave Opt Technol Lett 56:761–764, 2014; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.28187

Key words: filter monitoring; cavity resonators; adaptive filters; tunable circuits and devices; filters

1. INTRODUCTION
Tunable filters are the essence of emerging reconfigurable radios and spectrum-aware systems. Their capabilities of switching bands, changing communication standards, and handling jammers, among others, make them a very attractive choice for RF front-ends [1]. Yet, the flexibility of tunable filters comes at the cost of being potentially vulnerable to variations in terms of frequency drift caused by aging or environmental effects [2]. Also, tuning those filters may become a costly and time-consuming process.

This issue has been previously addressed using several techniques primarily focused on reducing the overhead cost and power consumption. The tuning technique in [3] relies on a two-step sequence of course and fine tuning that results in a frequency resolution of 160 Hz. However, an RF microprocessor and analog feedback loop is required, increasing the cost and size of the system. In addition, this technique takes several minutes to tune a cavity resonator, which might not be feasible for modern communication systems. In [4], fuzzy controllers automate the tuning process. In order to do that, frequency response information is required from each filter, requiring a costly network analyzer (NA) to tune the filters. A significant reduction in cost was presented in [5], where only a power detector and a frequency synthesizer are required along with a computer. Nonetheless, this technique demonstrates sequential tuning for each pole independently, which can be time consuming. Furthermore, this technique, as well as all the other techniques mentioned above, do not provide real-time measurement results, are costly, and consume a lot of power and volume, making them unsuitable as an embedded solution for tuning mobile-form factor systems.

This paper presents a new method for measuring the frequency response of each pole in a filter using simple circuitry that can be embedded in a system with reduced power consumption overhead. The measurement method is applied to an evanescent-mode cavity filter tuned by piezoelectric tuners [6]. Such filters have been proven to be programmable to provide multiple functionalities, which is essential in evolving spectrum-aware and frequency-agile systems [7]. The measurement method employs the natural response of a single-pole cavity resonator filter when excited by a current pulse as shown in Figure 1.

2. MEASUREMENT TECHNIQUE
When excited by a short current pulse, a single-pole filter (modeled as an RLC circuit) will produce a damped sinusoid as shown in Figure 1. The damping factor ($\zeta$) for a parallel RLC circuit is given by:

$$\zeta = \frac{L}{2R} \omega_0$$

where $L$ and $R$ are the inductance and the resistance in the RLC circuit, respectively, and $\omega_0$ is the frequency. From (1), it can be
observed that the rate of decay of the damped sinusoidal response is proportional to the resonant frequency. Consequently, the number of cycles with amplitude above a certain threshold decreases with increasing natural frequency of the resonator. As a result, frequency can be detected by counting the number of cycles (with amplitude above the threshold).

Since the amplitude of the damped sinusoid drops exponentially, it cannot be used to drive any ripple counter. Therefore, a limiting amplifier (HMC750LP4E) is used to restore the amplitude to emitter-coupled logic (ECL) compatible signal to successfully drive the 8-bit ripple counter (MC100E137). ECL is a current mode logic, which is used in high-speed wireline communication systems. Hence, it has been chosen due to its high-speed capabilities and off-the-shelf availability.

The sharp current pulse is generated by driving an npn transistor (BFP 740ESD) with a sharp voltage pulse, which comes from an XOR gate with a step function on one input, and a delayed version of that step function on the other input. The RC circuit of the series resistance (RDELAY) and the input capacitance of the XOR gate make the delay. RDELAY is set to 1 kΩ, giving a 1 ns current pulse. The resistor RLIM is used to limit the base current. The value of RLIM is 100 Ω. The schematic representation of the counting mechanism along with the pulse injection is shown in Figure 2.

The threshold at which the system stops counting cycles is set by biasing the differential inputs of the limiting amplifier 25 mV away from each other. If the difference was less than 25 mV, noise sources in the circuit would cause the limiting amplifier to generate random pulses, which would trigger the counter, giving false number of counts.

3. INTEGRATION WITH FILTER
In order to avoid any disturbance in the RF path of a filter from injecting a current pulse, a secondary cavity resonator is mounted on top of the pole to be measured in a filter. This way, the RF filter can still be used as none of its cavities are disturbed.

Any change in frequency on the cavity resonator of the filter will be inversely reflected on the secondary cavity resonator because the membranes’ movements induce opposite effects on frequency to the two attached cavities. The current pulse is applied on the secondary cavity resonator and the number of cycles is counted from its response. The stack up of the two cavities is shown in the inset in Figure 2. The membrane of the secondary cavity is coated with Parylene N to avoid shorting the bias voltage of the piezoelectric disk to ground.

4. RESULTS AND DISCUSSION
The presented measurement technique has been applied to a two-pole evanescent-mode cavity band-stop filter. The secondary cavity is mounted on one pole to trace it while it is being tuned. The two-pole filter is connected to a performance network analyzer (PNA) and the secondary cavity is connected to the current pulse–injecting circuit on one port, and the limiting amplifier and counter on the other port. This setup is show in Figure 3.

The control voltage (VCTRL) on the piezoelectric disk has been swept from 200 to −200 V and back to 200 V again to show the effect of hysteresis. The frequency of the pole in the filter changes from 1.422 to 1.65 GHz as shown in Figure 4.

As mentioned in the previous section, the frequency of the pole in the filter, \( f_{\text{pole}} \), is inversely proportional to the frequency in the secondary cavity \( f_{\text{sec}} \), or \( f_{\text{pole}} \propto 1/f_{\text{sec}} \). Also, from Section 2, the number of cycles the circuit counts, \( N \), is inversely proportional to \( f_{\text{sec}} \), or \( N \propto 1/f_{\text{sec}} \). As a result, the cycle-count reading
from the circuit is expected to be directly proportional to the frequency of the pole reading from the PNA, or \( (N \propto f_{\text{pole}}) \).

The frequency of the detected pole, obtained from the PNA, and the number of counts from the circuit have been recorded to show the relationship between reading from the circuit and actual operating frequency of the filter. The results are shown in Figure 5, where the direct monotonic relationship between the frequency of the pole in the filter and the number of cycles counted can be seen through the entire tuning range. The system has also been simulated in SPICE and the results are also shown in Figure 5. The error between the measurement and the simulation is mostly due to the SPICE model of the logic chips being unavailable. The model had to be constructed based on the information from the datasheet, hence the inaccuracy in the simulations.

Each point in Figure 5 is averaged 1024 times in order to suppress any noise and variations in readings. The averaging is done in an ALTERA DE2 FPGA board connected to the system. An ECL-to-TTL logic level converter (SY10H350) is used to interface the circuit to the FPGA. The secondary cavity is excited at a rate of 1.024 MHz. This means that one reading can be obtained in 1 ms.

As a result of the quantized output, there is a minimum change in frequency required to show a change at the output. The minimum frequency resolution \( f_{\text{min}} \) is defined by how much frequency change is required to cause a unity change in the counter (the inverse of the slope of the curve in Figure 5). This can be written as

\[
f_{\text{min}} = \frac{\Delta f_{\text{pole}}}{\Delta N}(2)
\]

Figure 6 shows the minimum resolvable frequency at each frequency range obtained from measurement and simulation. \( f_{\text{min}} \) can be as low as 2 MHz.

The measurement technique system runs at a 5 V power supply and consumes about 2.5 W. This amount of power could be reduced by duty cycling the circuit (turning the system ON during readout and turning it OFF when no readout is needed). When running the system at 1% duty cycle at 100 Hz, the power drops to 50 mW. Further power reduction can be achieved by implementing a dedicated chip.

5. CONCLUSION

In this paper, we present a new filter measurement technique to determine the operating frequency of a pole in a filter. This technique can be used to tune filters during operation, and to compensate for any frequency drift due to aging or environmental effect. The measurement technique relies on the number of cycles in the natural response of a cavity resonator, providing real-time results in a digital format with a resolution of down to 2–20 MHz (0.13–
1.3%). Using this technique can eliminate the need for expensive equipment for tuning, such as NAs, hence, saving on the cost of the calibration. The ability to measure and detect changes in frequency for each pole of a filter independently makes this technique much faster than conventional tuning methods. The technique is fully electronic and can be embedded in an adaptable communication system in real time.

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KOHCH SLOT LOOP ANTENNA FOR WIRELESS BODY-CENTRIC COMMUNICATION

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ABSTRACT: In this article, we describe the design of a coplanar waveguide (CPW) fed Koch slot loop antenna with a tuning stub. The antenna is optimized to be used in a wireless body-centric communication. The antenna is designed to operate in the ISM frequency band from 5.725 to 5.875 GHz (ITU-RS.138, 5.150, and 5.280). The study shows properties of the antenna, from the viewpoint of impedance matching, and radiation patterns. The impedance matching at the resonant frequency can be acquired by a tuning stub. The designed CPW-fed slot antenna exhibits a proper polarization, a low-radiation loss, and a weaker dispersion. We experimentally prove that the designed antenna efficiently excites surface waves and creeping waves. © 2014 Wiley Periodicals, Inc. Microwave Opt Technol Lett 56:764–766, 2014; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.28142

Key words: fractal antenna; Koch antenna; body centric communication; human body model

1. INTRODUCTION

At present, wireless body area networks (WBANs) attract attention, thanks to promising applications in the field or biomedical engineering, healthcare, assisted living, and so forth. In WBAN, the electric field intensity is requested to have a normal polarization with respect to a skin surface. If the whole body is requested to be covered by a transmitted signal, the antenna should show an omnidirectional radiation pattern in plane of a body surface. From the practical reasons, the antenna should be of a low-profile, a low-weight, and a compact size.

Recently, several papers have been published discussing the exploitation of body centric communication in health care, military applications, and other fields. For such applications, several types of antennas have been developed and tested. For example:

A miniaturized UWB stepped-slot antenna for medical diagnostic imaging was reported in [1]. A low-profile foam dielectric over-the-shoulder antenna based on coupled patches was described in [2].

In this article, we design a slot loop antenna. Considering duality theorem, the slot loop antenna can be understood as an equivalent of an electrical monopole, which is perpendicular to the surface of a body. To prolong the magnetic current line of the antenna, the shape of the loop follows the Koch fractal. The antenna structure is optimized to operate in vicinity of a human body in the ISM frequency band (5.725–5.875 GHz).

For simulations and optimization of the antenna located close to a body, we used a simplified three-layer chest model and a voxel Duke model [3]. The designed antenna is simulated and optimized in CST Microwave Studio. The impedance matching and the radiation pattern are main objectives of the design. The letter consists of five sections. In Section 2, we describe the designed antenna. Section 3 deals with a human body model. Section 4 compares simulation and measurement results. Section 5 concludes the article.

2. ANTENNA CONCEPT

To keep physical dimensions of the antenna constant when prolonging the magnetic current length of the loop, the fractal Koch geometry can be applied for shaping the loop. We follow the procedure of the iterative synthesis of the Koch fractal described in [4]. This procedure does not increase the dimension of the antenna, but the physical length of the wire is prolonged by a factor (4/3)n, where n is the number of the iteration.

The designed slot loop antenna is of a shape of a Koch snowflake created from the third iteration of the Koch fractal. The width of the slot is 0.40 mm. The length of each element is given by l_e = 0.5 \times m \times l_1 \times (4/3)^n, where m denotes the number of elements of the Koch snowflake, l_1 indicates the length of a straight wire, and n is the number of the iteration. The length of the element is chosen to be equal to the wavelength on the substrate.

In the top point, the slot of the Koch snowflake is interrupted by a narrow strip of the width w_2 = 0.50 mm. This interruption plays a role of a serial susceptance in the circuit of a magnetic current to be used for tuning the input impedance of the antenna in resonance (see Fig. 1).

The loop antenna is fed by a coplanar waveguide (CPW) with the characteristic impedance of 50 Ω. The width of the signal strip is A = 1.50 mm and the gap between the signal strip and the coplanar ground plane is B = 0.20 mm. We follow the procedure of the design of CPW feeder in [5]. The CPW feeder has a tuning stub of the length L_e = 3.04 mm and the width W_e = 1.26 mm. A proper impedance matching of the slot antenna can be achieved by setting the length and the width of the tuning stub [6]. The total size of the antenna is 25.30 × 16.00 mm².