A mechanically tunable GHz passive voltage element using microstrip resonator

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Abstract. The ability to detect very low RF signals can be used to wake-up a sensor node only when needed to sense or transmit data, reducing the sensor node power consumption. A passive voltage transforming stage from the antenna to a nano-watt receiver can be useful to reduce the minimum detectable RF signal. At high RF frequencies, the transformer is harder to implement owing to the low impedance of load capacitive loads. We demonstrate a gigahertz frequency range passive voltage transformer, using a mechanically tunable strip-line resonator at 1-2 GHz. The transformer exhibited a 19.5dB gain at 1.007 GHz and achieved gains over 19dB with load capacitances of 0.8-2.4pF. A piezoelectric AlN GHz transducer was also used to measure the gain of the resonator. An RF waveguide model has been developed which matches experiments well for both frequency and gain responses.

1. Introduction
An event-driven sensing system, awakened by external signals of relevant stimulus, can greatly reduce the power consumption compared to a system in which active power is consumed to amplify incoming signals. If the energy in the signals is used to trigger switches that do not consume power, nanowatt power consuming wake-up sensors can be implemented. Using nanowatt to microwatt energy harvesting systems, the ultralow power consumption sensor nodes can enable near perpetual function. The RF wake-up often requires switches that turn on at a minimum detectable RF signal, which can be increased through a passive transformer. A gain element is required to increase detectable RF voltage above the threshold of detection devices, while rejecting out-of-band signals. The gain of passive elements is limited by the ratio on the input to output impedance, and therefore requires high matching input impedance detectors, such as MEMS switches [1-2]. Prior works have demonstrated passive gain transformer filter at nanowatt power levels at MHz range [3-4]. Recently, a passive RF gain of 18.5dB was shown in the 400MHz range using LC tanks [5]. However, these devices are limited to MHz frequency range, as the gain is greatly reduced at higher frequencies because of the lower quality factor and lower input impedance at high frequencies. It is a significant challenge to achieve high transformer gains at GHz range, where many radio systems are present including cellular communications and Bluetooth bands.

To enable operation of sensor node at GHz frequency range, we developed a mechanically tunable passive voltage amplifier, which uses a microstrip line resonator for voltage amplification. A short in the transmission line was implemented using a copper clip to adjust the quarter wavelength resonator frequency. An equivalent circuit model was demonstrated to analyze the resonant behavior of the passive voltage transformer.
2. Resonator design and theoretical model

Figure 1 shows the GHz microstrip line quarter wave resonator, which has a copper bar folded clip that forms a short on the waveguide. At the quarter wavelength open-end, the current is minimum, and the voltage reaches its maximum value. Design parameters of the resonator are shown in Table 1, along with the attenuation in the dielectric material (Duroid 5880). The input voltage is coupled into the stripe line by adding the copper clip short boundary. Two magnets are used to clamp the clip to waveguide. This tunable boundary structure allows adjustments on the resonant frequency. The location of the boundary sets the effective length of the microstrip line to be \( l = l_0 - l_c \), where \( l_0 \) is the total length of the stripe line and \( l_c \) is the distance between the input terminal and the boundary. The resonant frequency is 
\[
    f = \frac{2\pi}{\lambda} = \frac{2\pi c}{4l_0 \sqrt{\varepsilon}}
\]

The approximate equality is coming from the impact of the coupling boundary which will be analyzed later in the equivalent circuit model. Figure 1c shows the resonant frequency of the system as a function of the location of the short boundary. It is shown that the resonant frequency of the system can be tuned in the order of hundred megahertz by changing the location of the coppered boundary within a few mm.

An equivalent circuit of the resonator is shown in Figure 2. A voltage source is cascaded the coupling boundary, a microstrip-line, and a capacitive load. The voltage source is associated with an output resistor. The copper clip boundary is modeled as an inductor \( L_{BC} \) and a parasitic resistor \( R_{BC} \), which

![Figure 1](image1.png)

**Figure 1.** (a) A schematic of the GHz resonator. (b) Top and side view of the resonator device. A fold copper clip is fixed with magnets that can move the shorted boundary condition. (c) Resonant frequency of the system as a function of the boundary position.

![Figure 2](image2.png)

**Figure 2.** The equivalent circuit model of the resonator system.

| Table 1. Resonator Design and Performance |
|------------------------------------------|
| **Microstrip line width** | 4.5mm | **Conductor loss** | 0.042 Np/m |
| **Microstrip line length** | 50.8mm | **Dielectric loss** | 0.0069Np/m |
| **Board thickness** | 1.575mm | **Total attenuation constant** | 0.0493Np/m |
| **Boundary inductance** | 1.01nH | **Unloaded Q factor** | 285 |

An equivalent circuit of the resonator is shown in figure 2. A voltage source is cascaded the coupling boundary, a microstrip-line, and a capacitive load. The voltage source is associated with an output resistor. The copper clip boundary is modeled as an inductor \( L_{BC} \) and a parasitic resistor \( R_{BC} \), which.
are connected in parallel with the microstrip line. A lossy transmission line is used to model the microstrip line. Loads are added at the end of the stripe line. The impedance of the load is expected to be high to match the large impedance of an open-end quarter wavelength transmission line.

The transmission (ABCD) matrix is used to characterize the cascade connections of two-port networks. ABCD matrix is defined in terms of the total voltages and currents as: 

\[
\begin{bmatrix}
Y_{in} \\
I_{in}
\end{bmatrix} = \begin{bmatrix}
A & B \\
C & D
\end{bmatrix}
\begin{bmatrix}
Y_{out} \\
I_{out}
\end{bmatrix},
\]

where \( \begin{bmatrix} A & B \\ C & D \end{bmatrix} \) is the total transmission matrix of the system (\( M_{Total} \)) and is obtained by multiplying the transmission matrix of each components. The transmission matrix of the source \( M_S \), boundary \( M_{BC} \), resonator \( M_R \) and load \( M_L \) are [6]:

\[
M_S = \begin{bmatrix}
1 & R_s \\
0 & 1
\end{bmatrix},
M_{BC} = \begin{bmatrix}
1 & \cosh(yd) \\
0 & \tanh(yd)
\end{bmatrix},
M_R = \begin{bmatrix}
\cosh(yd) & Z_0 \sinh(yd) \\
\tanh(yd) & \cosh(yd)
\end{bmatrix},
M_L = \begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix},
\]

where \( R_s \) is the source output impedance, \( y \) is the propagation constant, and \( d \) is the length of the microstrip line. The propagation of waves along the microstrip line is determined by \( y \), which depends on both the conductor loss \( \alpha_c \) and dielectric loss \( \alpha_d \). The loss performance of the resonator is listed in Table 1. The conductor loss is limited by the conductivity of the metal layer. Most metal has a conductivity of \( 10^{-7} - 10^{-8} S/m \), corresponding to a signal loss of \( 10^{-2} - 10^{-4} \) Np/m. The dielectric loss is determined by the dielectric material of the microstrip line. In this paper, we used the duroid 5880 board, which has a very low dielectric constant \( \varepsilon_r = 2.20 \pm 0.02 \). This low dissipation factor minimizes the dielectric loss in wave propagation and extends the performance of the resonator in GHz range. A dielectric loss of 0.0069Np/m was calculated for the duroid board, which is much smaller than the conductive loss. The cascaded networks are characterized by multiplying the transmission matrices of elements: 

\[ M_{Total} = M_S \times M_{BC} \times M_R \times M_L. \]

From the definition of the ABCD matrix, the voltage gain can be subtracted from the first element of the total transmission matrix:

\[
\frac{V_{out}}{V_{in}} = \frac{1}{A} = \frac{1}{M_{Total}[1,1]}.
\]

3. Simulation results and measurements of voltage gains

Based on the model developed above, the performance of the GHz resonator is simulated and compared with the experimental data in this session. Measurements are obtained with a GGB pico-probe (Model Number 34A), which has an input resistance of 10MΩ and an input capacitance of 0.1pF. The voltage gain is defined as the ratio of the output voltage obtained at the resonator end over the output voltage from a voltage source terminated by a 50Ω resistor.

Figure 3 shows the gain behavior of the GHz resonator without any load. A maximum gain of 19.5dB is obtained at 1.007GHz. Voltage gain from the equivalent circuit model is compared with the experimental data and fits well for both frequency and amplitude performance. Figure 4 shows the simulation of voltage gain with different source impedance. Higher amplification is found to be expected with smaller source impedance since more power could be coupled into the resonator with smaller resistance at the source side. The model predicts a voltage gain of 32.6dB with a 1Ω output impedance voltage source.

To test the impact of loads, a few picofarads capacitors were added as loads at the open-end of the resonators. Figure 5 shows the voltage gains measured by the pico-probe from a Network Analyzer. The voltage gains of the resonator remain above 19dB with load capacitances from 0 to 2.4pF. For a NEMS switch, typically with 100fF input capacitance, a voltage gain of ~19.4dB is expected. Although the resonant frequency is not independent of the load, a desired resonant frequency can be designed for specific loaded capacitance by varying the length of the microstrip line.
As a capacitive load representative, we used a recently developed GHz ultrasonic transducer as a load [7]. The transducer uses focusing transducers placed at one side of the wafer, which generates bulk acoustic waves through the silicon substrate adding in phase at the focus spot at the other side. An optical vibrometer was used to measure the vertical displacement of the focus. The effective area of the AlN transducer is $4770 \mu m^2$ and the AlN thickness is $2\mu m$. This corresponds to a capacitance of $1.9pF$. By adding the GHz resonator between the voltage source and the input side of the transducer, a higher displacement is expected with input actuation voltage remains the same. In figure 6, the transducer showed a 1.7 times higher displacement with the passive amplifier than without. The wire bond between the transducers and the resonator added parasitic impedance to the loads, which needs to be considered in resonator design. Better amplification is expected with resonant frequency matching and impedance matching.

**Figure 3.** Voltage gain from experiments and simulation.

**Figure 4.** Simulations on voltage gain with different source impedances.

**Figure 5.** Voltage gain with different load capacitance.
4. Conclusion
In this work, we developed a quarter-wave microstrip line resonator which can serve as a GHz passive voltage transformer. The developed device used a fold copper clip as the boundary condition to create the location of injection at the external voltage into the resonator. This structure allows for adjustments in resonant frequency. An equivalent circuit is proposed and can be used to analyze the voltage gain and resonant frequency of the GHz resonator. Load capacitances were used to testify the amplification of the transformer, which demonstrated gains over 19dB. The proposed passive amplifier can be used for GHz RF signal detection with high input impedance detector such as MEMS/NEMS switches.

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