A 60 GHz-band Coplanar MMIC Active Filter

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Abstract **— This paper presents the design and performance of a 60 GHz-band coplanar active filter. To compensate the loss of a passive filter, a resonator composed of a quarter-wavelength line is terminated by a negative resistance circuit. A cross-coupling is introduced to make attenuation poles at both sides of the pass-band. A fabricated filter with two resonators shows an insertion loss of 3.0 dB with a rejection of greater than 20 dB at 3 GHz-separation from a center frequency of 65.0 GHz. The size of the filter is 2.5 mm × 1.1 mm.**

I. INTRODUCTION

Increasing demands for high-speed multimedia links, such as wireless local area networks [1] and wireless home networks [2], stimulate the development of millimeter-wave transceiver. For these systems, filters are key components in order to suppress spurious signals out of the communication band sufficiently. Requirements for filters are a low insertion loss and high selectivity. At a microwave band, filters are widely implemented by active configuration because their loss issue can be solved and be small in size [3]-[5]. On the other hand, at millimeter-wave band, only one active filter operating at 32 GHz has been reported [6], because it is difficult to achieve a stable operation due to high sensitivity to the fabrication process.

In this paper, we present a 60-GHz-band active filter with two resonators terminated by negative resistances. We adopted a coplanar configuration because it not only eliminates backside process unlike a microstrip configuration, but also is suitable for a flip-chip technology [7][8]. We confirmed stable loss compensation due to the negative resistance for the fabricated filter. We also attained remarkable size reduction, compared to the passive filter with planar dielectric resonators [9] used in our transceiver module. To the best of our knowledge, this active filter is the one operating in the highest frequency band.

II. ACTIVE FILTER DESIGN

To realize active filters, many approaches have been reported [3]-[6]. In this work, we adopt a negative resistance approach, which is simplest and suitable for the millimeter-wave band.

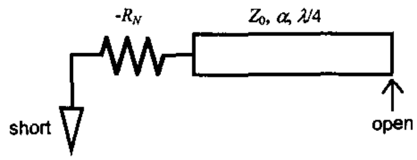

Fig. 1. Quarter-wavelength resonator terminated by a negative resistance.

A. Loss Compensation

Fig. 1 shows the principle of loss compensation due to a negative resistance for a resonator. The resonator consists of a quarter-wavelength line $\lambda/4$. Each port of the resonator terminates in open- and short-circuits. The Q-factor of the resonator is limited due to ohmic, dielectric, and radiation losses. In order to compensate the loss, the short-circuited terminal is replaced by a negative resistance $-R_N$. When the line has a characteristic impedance Z_0 and an attenuation constant α, the loss L due to a round path of the resonator is expressed as equation (1). The gain Γ produced by the negative resistance is expressed as equation (2). When $L \times \Gamma = 1$ is satisfied, the loss of the resonator is completely compensated and the Q-factor is infinite. In this case, the negative resistance should satisfy equation (3). When the negative resistance is larger than that given by equation (3), the resonator has a loop-gain and oscillation may occur.

$$L = e^{-\lambda\alpha/2} \tag{1}$$

$$\Gamma = \frac{-R_N - Z_0}{-R_N + Z_0} \tag{2}$$

$$R_N = \frac{Z_0\left(e^{\lambda\alpha/2} - 1\right)}{e^{\lambda\alpha/2} + 1} \tag{3}$$

B. Negative Resistance Circuit

A negative resistance not only satisfies equation (3), but also is constant over a wide frequency range in order to compensate the loss of a resonator without instability and oscillation. Fig. 2 shows the schematic of a negative resistance circuit using an FET as an active device. A negative resistance is usually obtained by connecting a capacitive line to the source terminal at higher frequencies than 10 GHz. The capacitive line consists of a short-

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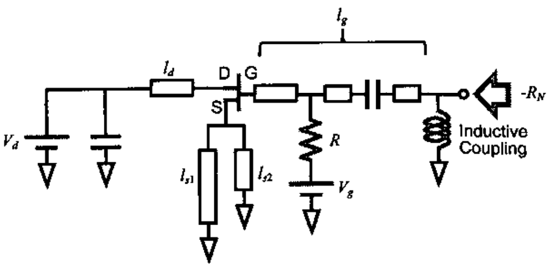


Fig. 2. Schematic of the negative resistance circuit.

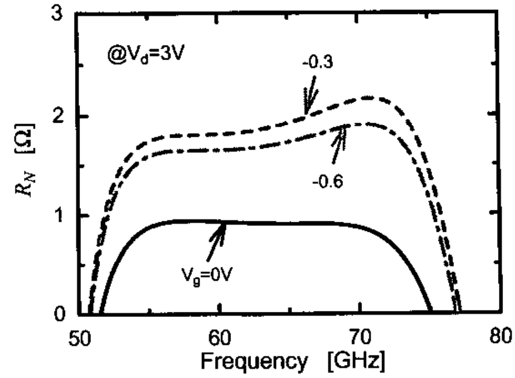
circuited line l_{s1}, which is longer than a quarter-wavelength. In this work, we connected another short-circuited line l_{s2} in parallel, which is shorter than the capacitive line. The additional line provides a constant negative resistance over a wide frequency band. In our configuration, we place the output terminal on the gate terminal side, because it can eliminate a large drain bias network. The gate voltage V_g is applied via a resistance with a large value R of 3 KΩ. The negative resistance is inductively coupled to an outside resonator. Its value is tuned by the inductance.

We employed a coplanar waveguide with a ground-to-ground distance of 70 μm for the line. Its attenuation constant α is estimated to be 0.039 mm^{-1} by using an EM-simulator. The required value of a negative resistance -R_N is derived to be about -1 Ω from equation (3). However, the actually required value should be larger than –1 Ω in a negative direction considering additional losses at I/O and inter-stage connections.

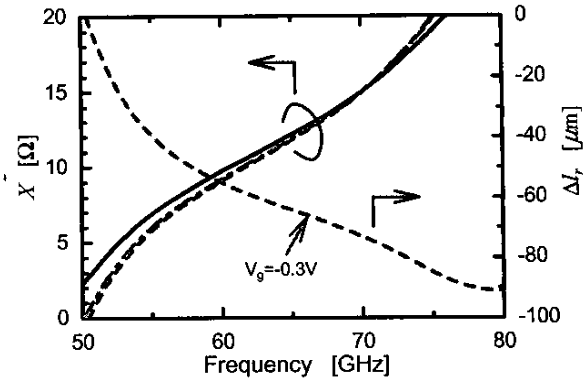
Fig. 3 shows a calculated impedance of the negative resistance circuit after optimization of the length l_d, l_g, l_{s1}, and l_{s2}. The resistance part -R_N has a constant value of about –2 Ω at V_g of –0.3 V from 55 to 70 GHz. The figure indicates that its value can be tuned by the gate voltage with a constant value over this frequency range. The reactance part X has a non-zero value. However, the small reactance can be compensated by changing the length of the resonator as $\lambda/4+\Delta l_r$, where Δl_r is derived from equation (4) using the phase constant β of the coplanar waveguide.

$$\Delta l_r = -\frac{1}{2\beta}\tan^{-1}\frac{2X}{Z_0} \tag{4}$$

In Fig. 3(b), the length Δl_r at V_g of –0.3 V is also shown. The length Δl_r is not constant over the frequency band, but increases in a negative direction with increasing frequency. As a result, the compensation using a specific value (e.g. –55 μm at 60 GHz) decreases the bandwidth of the filter, compared to the case that the length Δl_r is constant over the frequency band. It helps the design of a narrow bandwidth filter. The optimized parameters were l_d = 200 μm, l_g = 450 μm, l_{s1} = 700 μm, and l_{s2} = 350 μm.



(a)



(b)

Fig. 3. Calculated impedance of the negative resistance circuit: (a) resistance part; (b) reactance part.

C. Active Filter Circuit Diagram

Fig. 4 shows the circuit diagram of an active filter. We adopted a two-stage filter in order to attain a flat band with the minimum number of stages. Capacitive couplings were used for I/O and inter-stage connections. We also introduced a cross-coupling to improve the skirt characteristic of the filter. For this purpose, an inductive coupling is incorporated between I/O ports to make attenuation poles at both sides of the pass-band.

III. Experimental Results

Fig. 5 shows the chip photograph of a fabricated active filter MMIC. The chip size is 2.5 mm × 1.1 mm × 0.15 mm. The active devise is an AlGaAs/InGaAs heterojunction FET (HJFET) with a gate-length of 0.15 μm and a gate-width of 50 μm × 2. Different voltages for

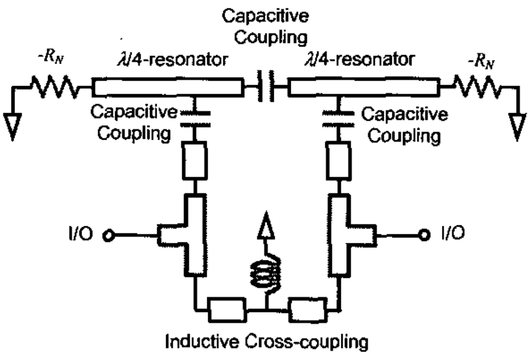


Fig. 4. Schematic of the active filter with two resonators.

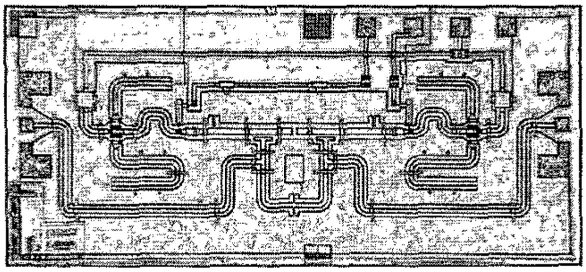


Fig. 5. Chip photograph of the active filter MMIC (2.5 mm × 1.1 mm × 0.15 mm).

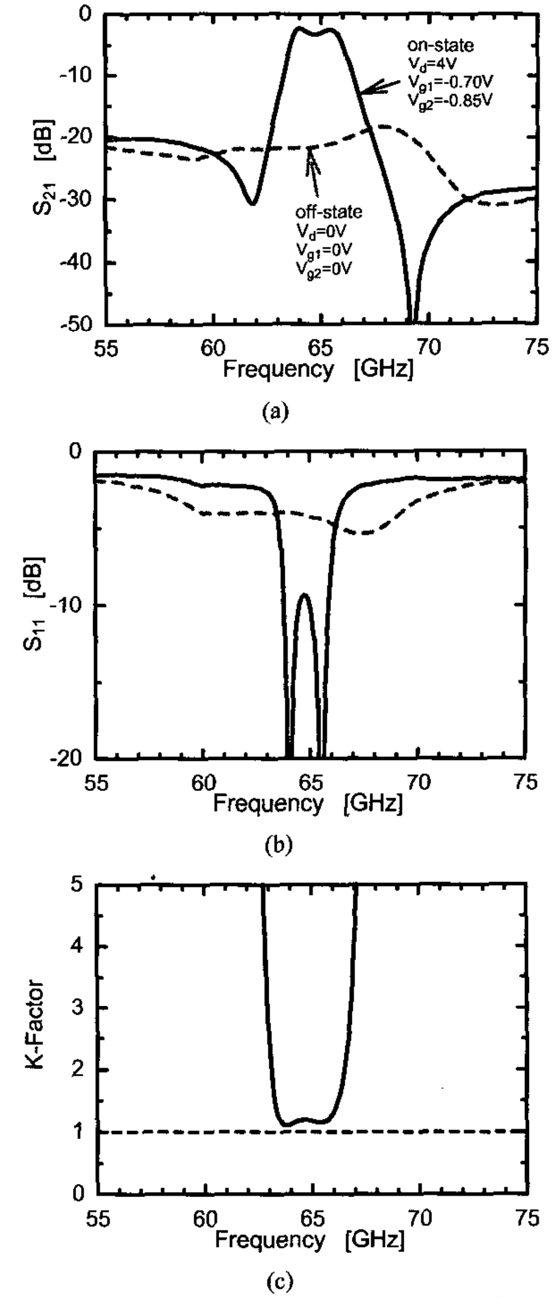


Fig. 6. Measured filter performance of the active filter: (a) transmission characteristics; (b) matching characteristics; (c) K-factor.

each HJFET can be applied to compensate the difference between two negative resistances. We adopted a wider ground-to-ground distance of 70 μm for coplanar resonators than 52 μm for other lines to increase the Q-factor of resonators. For the design, we calculated S-parameters of passive components such as I/O and inter-stage connections and a T-junction by an EM-simulator.

Fig. 6 shows the measured small-signal characteristics of the active filter. The biases were determined to satisfy the condition that the insertion loss was smallest with the K-factor of larger than 1. Attenuation poles due to a cross-coupling were clearly observed at both sides of the pass-band and improved the skirt characteristics near the pass-band. The measured insertion loss was 3.0 dB at the center frequency of 65.0 GHz with a 3-dB bandwidth of 2.6 GHz. The rejection of larger than 20 dB was obtained at a 3 GHz-separation from the center frequency. Considering the feed loss, which corresponds to the return loss in the stop-band, the intrinsic insertion loss of the filter was estimated at better than 1.5 dB. The on/off ratio of 15 dB was obtained. We also confirmed that instability and oscillation did not occur during the measurement. It indicates that the negative resistance circuit had a stably constant value over this frequency band.

IV. CONCLUSION

An active filter MMIC with two resonators was developed for 60 GHz-band applications. The loss of a resonator was compensated successfully using a negative resistance. A cross-coupling was introduced to improve the stop-band rejection of the filter. The fabricated filter exhibited a low insertion loss and steep skirt characteristics without instability. This active filter is promising for applications to high-speed wireless communication systems.

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