Enhanced-Q Microstrip Bandpass Filter with Coupled Negative Resistors

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Abstract - Microstrip technology offers a very compact realization of a microwave filter. Unfortunately, the losses in normal thin-films are usually so excessive that narrow-band microstrip filters are impractical. However, high-Q microstrip filters can be realized by augmenting them with active elements. This paper presents a microstrip bandpass filter with negative resistors coupled to each resonator. The negative resistors negate the losses in each resonator and yield an enhanced-Q. A new theoretical framework for this Q-enhancement technique and a design methodology are presented. The simulated and measured performance of the filter are given.

I. Introduction

Traditionally, microwave filters have been implemented as passive networks of transmission-line, and as discrete lumped-elements. Consequently, microwave filters remain quite bulky, especially in comparison to microwave monolithic integrated circuits (MMICs). The current trend is to reduce the mass and size of communications, radar, and signal processing systems. MMICs have made it possible to significantly reduce the size of amplifiers, and other non-filtering circuits, without performance degradation. Passive filters are generally not amenable to miniaturization without an appreciable trade-off in performance. Usually, the best filters are built using waveguide. Filters built using microstrip or lumped-elements are smaller, but always result in higher loss and less isolation, for a given filter order. While SAW filters offer the best isolation-to-size ratio, they are extremely lossy. In addition, SAW technology is currently limited to the low microwave frequency range (typically less than 2 GHz). Recent advances in HTS technology have allowed filter size reduction while maintaining or exceeding the required performance. However, the high cost of cryogenic systems may prove to restrict the application of HTS technology to a small number of niche applications. The search for a solution to microwave filter size reduction, without sacrificing performance, has led investigators to consider active circuit implementations [1-5].

Microstrip technology offers a very compact realization of a microwave filter. Unfortunately, the losses in normal thin-films are usually so excessive that narrow-band microstrip filters are impractical. However, high-Q microstrip filters can be realized by augmenting them with active elements. This paper presents an active microstrip bandpass filter. Q-enhancement is accomplished with negative resistors coupled to each resonator. A number of other authors have also used negative resistance loss compensation techniques [1-3]. In particular, Chang and Itoh [3], have reported a method which utilizes coupled negative resistance to provide loss compensation. They used this technique to design end-coupled half-wave resonator microstrip filters. Loss compensation was provided by a quarter-wave section of line coupled to each of the half-wave resonators. We also use coupled negative resistors to provide Q-enhancement of a microstrip filter, but on a high-dielectric constant material ($\varepsilon_r = 24$). In addition, we present a new theoretical framework for the coupled negative resistance Q-enhancement technique, and a design methodology. The method is used to design an active hairpin microstrip filter. The simulated and measured performance of the filter are presented.

II. Method of Q-Enhancement

Consider the lossless first-order BPF given in Figure 1a. This circuit is a series resonant structure with a couple of external terminating impedances (R_T). The finite Q of the capacitor and inductor can be modeled by a single series resistor. The same filter, with loss, can be modeled by the circuit given in Figure 1b, where R_o represents the combined losses of C_o and L_o. A method for compensating the losses represented by R_o is to replace L_o with a transformer whose secondary winding is terminated in a negative resistance, as shown in Figure Figure 1c. [3] Here $R_n < 0$, L_1 and L_2 are the inductances of the primary

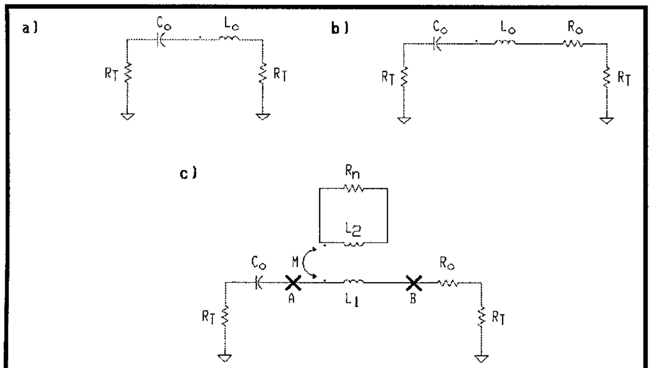

Figure 1. First order BPF a) lossless, b) lossy, c) with loss compensation

and secondary windings respectively, and M represents the mutual inductance between the two inductors. The problem is to select L_1, L_2, M and R_n such that the impedance across the nodes A and B is,

$$Z_{AB}(\omega) = R_{AB}(\omega) + j\omega L_{AB}(\omega) = -R_o + j\,\omega\, L_o \quad (1)$$

In this way, the losses Ro are canceled and the resonant frequency remains unaffected. We can show that equation 1 is satisfied when,

$$R_o = R_{AB}(\omega) = \frac{(\omega k)^2 L_1 \omega_x}{\omega_x^2 + \omega^2} \quad (2)$$

$$L_o = L_{AB}(\omega, k) = L_1\left(1 - \frac{(\omega\, k)^2}{\omega_x^2 + \omega^2}\right) \quad (3)$$

where,

$$\omega_x = \left|\frac{R_n}{L_2}\right| \quad (4)$$

and k is the transformer's coupling coefficient defined as,

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (5)$$

Equations 2 and 3 must be satisfied across the entire filter bandwidth in order to achieve a lossless filter with no perturbation to the frequency response. However, both of these expressions are functions of frequency. This means that these conditions cannot be exactly satisfied over the entire filter bandwidth. This will result in some perturbation to the ideal filter performance. The parameters L_1, L_2, k, and R_n must be selected to minimize the performance variation across the filter band

$R_{AB}(\omega)$ has a second-order highpass shape with cutoff frequency equal to ω_x. Figure 2 is a plot of $R_{AB}(\omega)$ normalized to $k^2\omega_x L_1$. This plot clearly shows that for small variation of $R_{AB}(\omega)$ across the filter band, the following condition must be met:

$$\omega_o >> \omega_x \quad (6)$$

where ω_o is the center frequency of the filter.

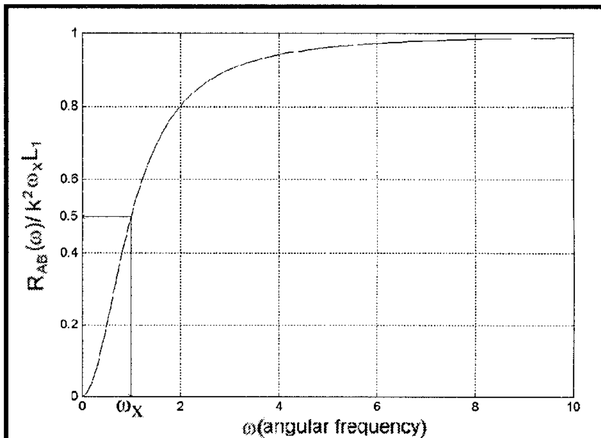

Figure 2. Plot of $R_{AB}(\omega)$ normalized to $k^2\omega_x L_1$. This plot clearly shows that $R_{AB}(\omega)$ has a 2nd-order highpass shape with cutoff frequency $\omega_x = |R_n/L_2|$.

L_{AB} is plotted in Figure 3 normalized to L_1 as a function of both frequency and the transformer coupling coefficient (k). The plot shows that for low variation in $L_{AB}(\omega,k)$ with frequency, equation 6 must again be satisfied. In addition, the variability in $L_{AB}(\omega,k)$ with frequency decreases with decreasing k.

The transformer coupling coefficient (k) and ω_x cannot be selected independently of each other. Solving equation 2 and 3 for k at $\omega = \omega_o$ yields,

$$k = \sqrt{\frac{1+\alpha^2}{1+\alpha\dfrac{\omega_o L_o}{R_o}}} \quad (7)$$

$$\alpha = \frac{\omega_x}{\omega_o} \quad (8)$$

which clearly shows the inter-relationship between ω_x and k. In practice, values of α within the range $0.1 \le \alpha \le 0.25$ result in practical and small values for k, and also satisfy equation 6.

Having determined k and ω_x, L_1 is determined at ω =

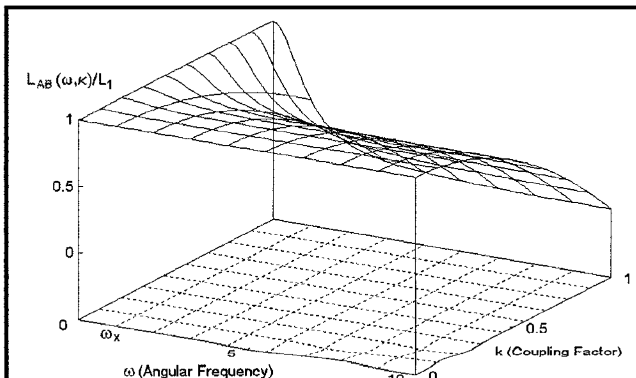

Figure 3. Plot of $L_{AB}(\omega,k)/L_1$ which shows that for minimum variation of $L_{AB}(\omega,k)$ with frequency, equation 6 must hold and k must be small.

ω_o from equation 3. Finally, L_2 is determined by selecting an R_n which yields a reasonable inductance in equation 4. Alternatively, we could set $L_2 = L_1$ and determine R_n from equation 4.

In summary, loss compensation is achieved by replacing the inductor L_o in each resonator with a transformer whose secondary is terminated in a negative resistance. The transformer's parameters, and the value of negative resistance required, are determined as follows:

STEP 1: Estimate R_o from the expected Q of the inductors and capacitors, or from the measured loss of a representative structure.

STEP 2: Select ω_x such that equation 6 is satisfied (i.e. $\omega_x << \omega_o$), and such that the value of k, given by equation 7, is relatively small and is realizable.

STEP 3: Determine L_1 from equation 3.

STEP 4: Set $L_2 = L_1$ and determine R_n from equation 4.

III. MICROSTRIP ACTIVE FILTER DESIGN

A lossless 4 GHz third-order bandpass filter with a 0.01 dB equal-ripple bandwidth of 72 MHz was established as the design goal. The first step in the design process was to select a suitable passive lumped-element BPF prototype. We assumed infinite Q for all the elements in arriving at the passive filter prototype. We designed, built, and tested a microstrip BPF, which has a similar topology as the final active filter, to obtain an accurate estimate of the losses. The losses were then modeled with a resistor R_o in series with each of the inductors. The method described in section II was used to introduce loss compensation to the filter prototype. Each of the inductors in the passive prototype were subsequently replaced with the coupled transformer and negative resistor Q-enhancement network. The resulting lumped-element prototype filter, with loss compensation, is given in Figure 4

The lumped-element prototype filter was translated into a microstrip equivalent network. A combination of electromagnetic simulation and linear circuit simulation was used to arrive at the final geometry. The transformers were realized with small lengths of coupled microstrip lines. The lengths and widths of the coupled lines are set to give the required inductances L_1 and L_2 The coupled-lines do not necessarily need to be a quarter-wavelength. The microstrip coupler does not behave like an ideal transformer, but reasonable correspondence can be achieved over a narrow bandwidth.

The negative resistor was built using a Fujitsu FHR10X HEMT. The negative resistance is produced at the gate terminal of the HEMT by terminating the drain

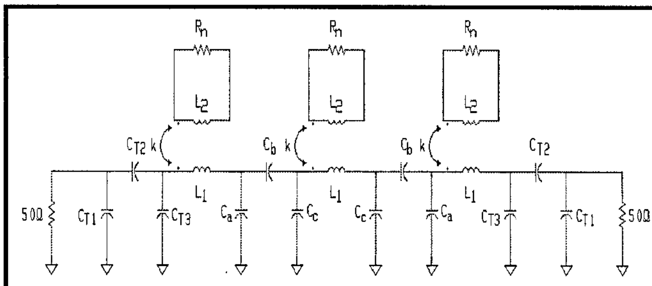

Figure 4. Active filter prototype.

with the appropriate impedance. The gate matching circuit reduces the imaginary component of the gate impedance to zero and transforms the negative resistance to the desired level.

The negative resistor is connected across the secondary of the microstrip coupled-line transformer by grounding one of the coupled-ports with a quarter-wave open-circuited radial stub, and by connecting the other coupled-port to the negative resistor (see Figure 5).

Initially, the active negative resistor was designed to provide the resistance calculated using the method described in section II. Subsequently, the active resistor matching circuits were re-optimized until an optimum filter performance was achieved. This was necessary because the microstrip components do not behave like ideal lumped-elements, as assumed in the analysis of section II.

The final filter layout is given in Figure 5. A 0.020 inch thick $LaAlO_3$ substrate was used for the microstrip components. The individual circuits are mounted on a single THERMKON carrier using silver-loaded epoxy. The overall footprint of the filter is 0.652 in. X 1.365 in.

The simulated filter performance is given in Figure 6. The plots clearly indicate the lossless characteristics of this filter. The simulation was performed by using an electromagnetic simulator to obtain the S-parameters of each of the individual passive microstrip circuits. The overall filter performance was then obtained by cascading the circuit S-parameters with the HEMT S-parameters.

IV. MEASURED FILTER PERFORMANCE

Test data for the active filter was obtained using an HP8720 vector network analyzer and an in-house microstrip test-jig. The HEMTs were biased at a drain current of approximately 5 mA and $V_{DS} = 2.0$ V. The measured performance of the filter is given in Figure 6 superimposed on the simulated performance. The filter has a measured center frequency of 3900 MHz which is slightly lower than the predicted center frequency (3923 MHz). The measured filter equal-ripple bandwidth is 37.5 MHz. Port 1 of the filter has a measured return loss

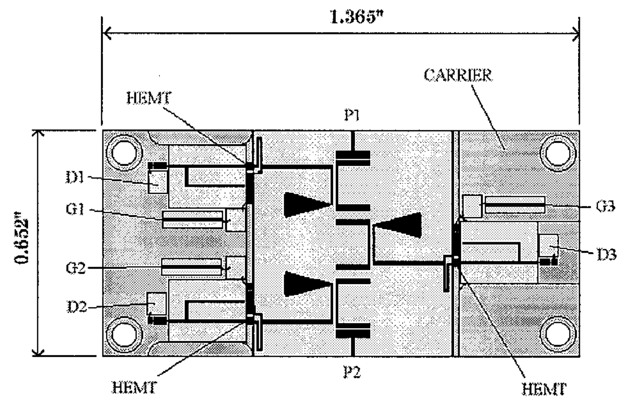

Figure 5. Final layout of the active filter. P1 and P2 are the RF ports. Bias is applied to the pins labeled D1, D2, D3, G1, G2, and G3.

of 15 dB whereas the return loss at port 2 is only 10 dB. An interesting feature of the active filter is that the passband ripple is very good (0.075 dB_{pp}), despite the poor VSWR. The filter has a center frequency loss of 0.05 dB which is very close to the simulated result. The filter provides in excess of 35 dB isolation at 140 MHz offsets both above and below the center frequency. The simulated isolation agrees relatively well with this measurement.

The effect of the negative resistors on the filter performance is clearly displayed in Figure 7. This figure shows the measured performance of the filter with the HEMTs biased on and off. The Q improvement with the devises biased-on is clearly evident from these plots. The filter has 15 dB center frequency loss when the HEMTs are turned off.

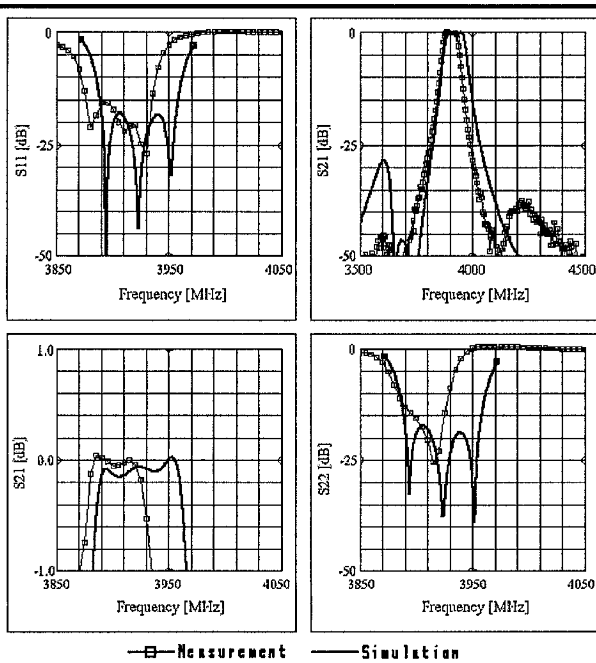

Figure 6. Measured and simulated performance of the active filter.

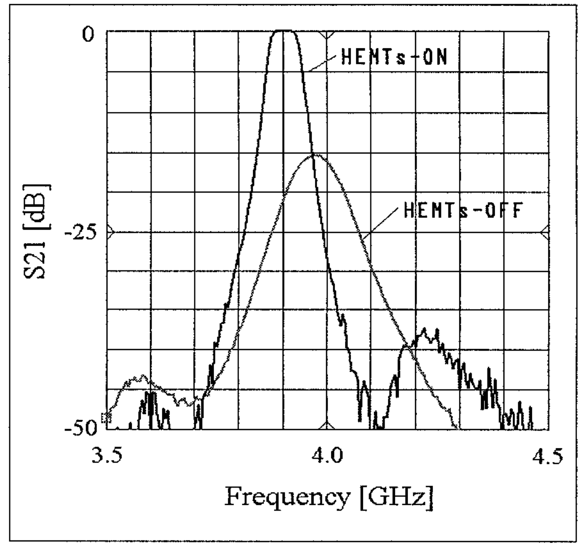

Figure 7 Measured performance of the active filter with the HEMTs biased-on and off.

V. Conclusions

This paper describes the development of an active microstrip filter using coupled negative resistors for Q-enhancement of each resonator. A detailed description of the filter design and measured performance of the filter are presented. A new theoretical description for the coupled negative resistance Q-enhancement technique, and a filter design methodology are given. The theory provides a useful method for the initial design of the coupled negative resistance class of active microwave filters. CAD tools can then be used to arrive at the final filter design.

Acknowledgment

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