MMIC Active Bandpass Filters Using Varactor-Tuned Negative Resistance Elements

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***Abstract*—This paper describes techniques for realizing microwave active filters using single transistor active resonators in a negative resistance configuration. The negative resistance topologies for both bipolar (AlGaAs/GaAs HBT) and FET (MESFET or HEMT) devices are studied and compared. The essence of the technique is that the input reactance of the transistor circuit resonates with an external capacitor or inductor, whilst the negative resistance is used to compensate for the losses in the resonator. It is shown that the FET device is ideally suited for this application as it can have a varactor-controlled negative resistance component. Three-stage and two-stage monolithic varactor-tuned bandpass filters have been demonstrated using this technique. The measured response of the three-stage filter exhibits a 120 MHz 3 dB-bandwidth centered on 2.3 GHz, 0 dB insertion loss with only ±0.1 dB ripple in the pass-band, up to 100 dB of stop-band attenuation at low frequencies, and over 50 dB of rejection up to 6 GHz. The two-stage filter exhibits a 400 MHz 3 dB-bandwidth centered on 4.7 GHz, with tunable insertion gain and only ±0.1 dB ripple in the pass-band.**

I. Introduction

FOR FULLY INTEGRATED microwave subsystems there is considerable interest in active filter techniques for MMIC realization. Conventional active filter techniques, widely used at low frequencies for integrated circuits, are not directly applicable to microwave filters because of the lack of suitable Op-amps at such high frequencies. A number of alternative techniques have been reported in the literature for microwave active filters. Sussman-Fort [1] has presented a second-order bandpass filter using actively-coupled resonators, and Bonetti [2], [3] has applied a similar technique to the design of wideband MMIC active filters and has achieved very high performance. As an alternative approach, the Q of a resonator can be increased by amplifying the signal within the resonator either with an active loop or by coupling negative resistance into the resonator. These techniques have been applied to dielectric resonators [4], and to microstrip resonators [5], [6]. Itoh achieved a 200 MHz wide bandpass filter centered on 10.5 GHz, with approximately 20 dB rejection 200 MHz from the band edges. In addition, tuning of the filter using varactors and using optical control was demonstrated [7], [8]. Transversal and recursive filter techniques have been taken from the digital filter domain and applied to microwave filters by a number of researchers. Jutzi [9] gives an extensive account of transversal filters realized with a distributed amplifier type topology. Rauscher also published some important work on microwave active filters using transversal and recursive techniques [10]. More recently, Schindler and Tajima [11] have used these techniques in monolithic filters with excellent results. Lastly, active filters can be designed using novel types of active resonator. For example, the active inductor [12]–[14] can be used as a substitute for spiral inductors in a lumped *L-C* filter, and other designs of high performance active resonator can be devised [15].

While active filters have the major advantages of small size with high selectivity, their limitations in terms of the DC power required, as well as the sensitivity to fabrication tolerances and environmental conditions, make it essential that active filter topologies which are simple and yet fully tuneable are identified for MMIC realization. In this paper, an active filter technique is described which is based on using an FET negative resistance circuit to realize a new type of lossless active resonator. Active resonators have obvious advantages of small size and high Q-factor for monolithic filters. However, active resonators previously reported for MMIC filters have tended to use fairly complex topologies which require many power supply lines for different settings of transistor and varactor bias voltages. For example, FET-based lossless active inductors require at least two FET's and may in practice need as many as four to achieve zero loss and good tunability. In contrast, in three particularly notable MTT papers [16]–[18] simple active inductor resonators have previously been demonstrated at UHF frequencies by using single bipolar transistors connected in a common-base configuration. In this technique the circuit can be designed to provide negative resistance to compensate for filter losses.

It has been shown before that MESFET's can also be used as negative resistance elements to compensate for losses in microwave active filters [5], [6]. It can be considered that the device is operating in a reflection-type amplifier mode in this case, amplifying the signal within the microstrip resonator. However, in this paper we use the FET negative resistance circuit as an integral part of a lumped *L-C* resonator, in a similar manner to the previous UHF filters employing bipolar transistors. When comparing the use of bipolar and FET devices in this technique it is found that they must be used in a fundamentally different way, and whilst the negative resistance principle is similar the actual filter topology becomes quite distinct. We explain this difference by first

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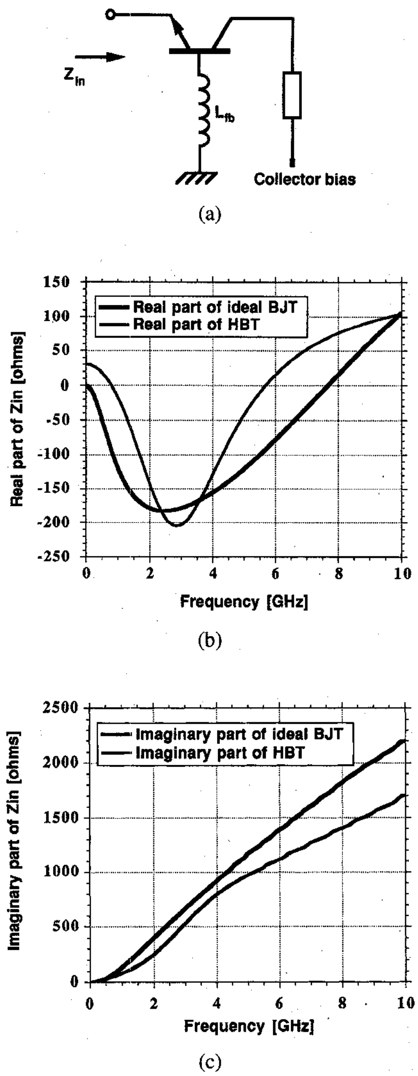


Fig. 1. Bipolar transistor negative resistance circuit. (a) Topology, (b) input resistance for an ideal bipolar transistor compared with a practical HBT, (c) input reactance for an ideal bipolar transistor compared with a practical HBT.

looking at the basic characteristics of the bipolar and FET negative resistance circuits.

II. Negative Resistance Circuits

In this section we will analyze and compare the different negative resistance circuit topologies used for bipolar and FET devices before describing the use of each type in forming lossless lumped resonators. The negative resistance topologies are mostly used in reflection type oscillators [19]. In the case of the bipolar transistor (BJT or HBT) the device is connected in common-base configuration with an inductive feedback element, as shown in Fig. 1(a), and the negative resistance is developed at the emitter terminal. In the case of the FET (MESFET or HEMT) the device can be in common-gate configuration with an inductive feedback element or in common-source configuration with a capacitive feedback element, as shown in Fig. 2(a), (b), and (c). The circuits provide feedback by circulating the output current through the feedback element in such a way that the voltage is developed across the feedback element becomes part of the input signal.

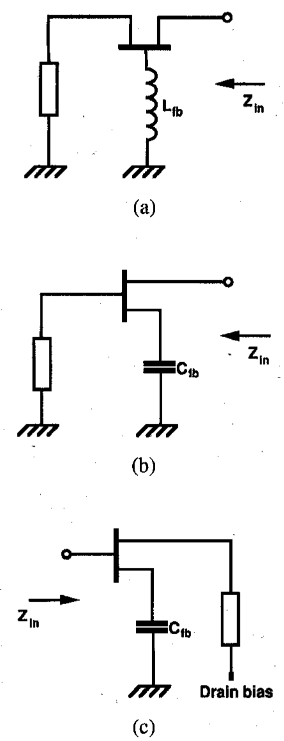


Fig. 2. FET negative resistance circuit topologies. (a) Common-gate with inductive feedback, (b) common-source with capacitive feedback and drain output, (c) common-source with capacitive feedback and gate output.

In the case of the BJT, the input impedance at the emitter terminal is given by [17]

$$Z_{\text{in}} = (1-\alpha)Z_b + Z_e. \tag{1}$$

Where Z_b is the impedance of the base feedback element, Z_e is the impedance of the base-emitter junction (R_{be} and C_{be} in parallel), and α is given by

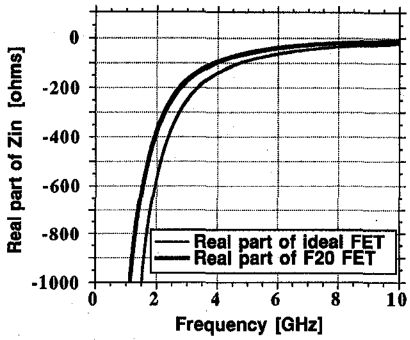
$$\alpha(f) = \alpha_0 \frac{e^{-j2\pi fT}}{1 + j\frac{f}{f_\alpha}}. \tag{2}$$

Where α_0 is the DC forward current-transfer ratio, T, is the associated time delay, and f_α is the 3 dB cut-off frequency for α. When the feedback element Z_b is an inductance L, it can be shown that the input resistance and reactance are as follows

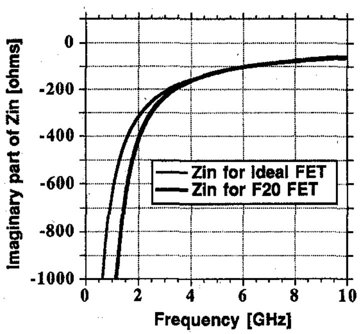
$$\text{Re}[Z_{\text{in}}] = -2\pi f L \frac{\left(\alpha_0 \sin 2\pi fT + \frac{f}{f_\alpha}\alpha_0 \cos 2\pi fT\right)}{1 + \left(\frac{f}{f_\alpha}\right)^2} + \text{Re}[Z_e] \tag{3a}$$

$$\text{Im}[Z_{\text{in}}] = 2\pi f L \frac{\left(1 + \left(\frac{f}{f_\alpha}\right)^2 - \alpha_0 \cos 2\pi fT + \frac{f}{f_\alpha}\alpha_0 \sin 2\pi fT\right)}{1 + \left(\frac{f}{f_\alpha}\right)^2} + \text{Im}[Z_e]. \tag{3b}$$

The device presents a negative resistance in series with an inductance. It is important to note that this behavior relies on the high frequency phase shift in the α of the transistor. With a



(a)



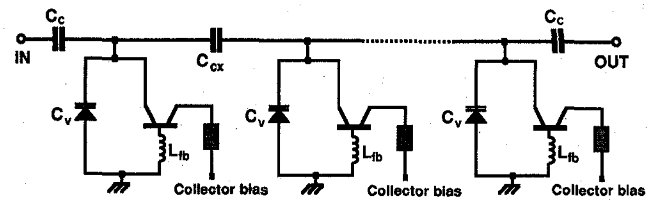
(b)

Fig. 3. (a) Input resistance for an ideal FET compared with an F20 FET, (b) input reactance for an ideal FET compared with an F20 FET.

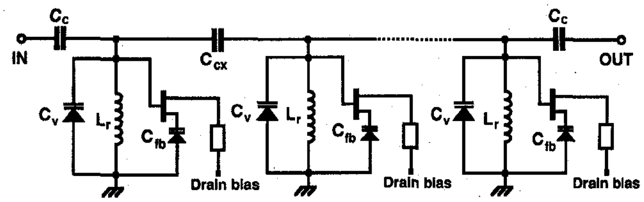
β or transconductance model we find that the subtle difference in the way the delay is incorporated makes a significant difference; negative resistance is still achieved but a very much larger base inductance is required. In Fig. 1 the resistance and reactance calculated from the above expressions is compared with a simulation using an AlGaAs/GaAs HBT fabricated at King's College in the Physical Electronics Research Group. A feature of most HBT's is a high base resistance, and because of this and the other parasitics in the real transistor it is not possible to achieve negative resistance without introducing an inductance in the collector. From Fig. 1, however, it is apparent that the above expressions give a good indication of the negative resistance achievable from the device.

Hopf *et al.* successfully used the common-gate FET negative resistance circuit topology of Fig. 2(a) to realize an L-band active filter in coplanar waveguide [20]. A common-source topology with a capacitive feedback element is a well-known alternative which has been used for lossless resonator circuits [21], [22]. The common-source topologies of Fig. 2(b) and (c) have the advantage that the feedback capacitance can be implemented as a varactor diode, providing precise control of the negative resistance value. In the filters presented here we have used the topology of Fig. 2(c) because we have found that it gives the required values of negative resistance even when the FET parasitic resistances and capacitances are taken into account. In the ideal case, the FET negative resistance circuit of Fig. 2(c) has an input impedance at the gate terminal given by [6]

$$Z_{\text{in}} = -\frac{g_m}{\omega^2 C_{gs} C_{fb}} - j\left(\frac{1}{\omega C_{gs}} + \frac{1}{\omega C_{fb}}\right). \tag{4}$$



(a)



(b)

Fig. 4. (a) Bandpass filter topology using bipolar transistor negative resistance elements. (b) Varactor tuned bandpass filter topology using FET negative resistance elements.

This input impedance comprises a negative resistance in series with C_{gs} and C_{fb}. Fig. 3 compares the resistance and reactance calculated from this ideal FET model with that achieved with a GEC-Marconi Foundry MESFET. Again, the simple model gives an accurate picture of the behavior of the real device, although in practice the negative resistance can be considerably increased at high frequencies by employing a reactive drain termination.

III. The Active Filter Topologies

In the case of the bipolar transistor the input impedance at the emitter is inductive and so the lossless resonator is formed by adding a capacitor in parallel with the transistor. Fig. 4(a) shows a multiresonator filter circuit employing this technique. In contrast, the FET circuit of Fig. 2(c) has a capacitive input impedance and so an inductor must be added in parallel to create the lossless active resonator. Fig. 4(b) shows the circuit diagram of a multi-resonator filter topology employing this new technique. Fig. 5 shows the simplified equivalent circuit of a single FET active resonator. The input impedance of the active resonator is as follows

$$Z_r = \frac{(j\omega L + R_{\text{ind}})(\omega C Z_{\text{neg}} - j)}{j(\omega^2 LC - 1) + \omega C R_{\text{ind}} + \omega C Z_{\text{neg}}}. \tag{5}$$

Where, R_{ind} = Parasitic resistance of the inductor, and Z_{neg} = Input impedance of the negative resistance circuit.

Note that varactor diodes have been incorporated as the source feedback elements and in parallel with the resonators. The latter varactors have the obvious function of providing frequency tuning of the individual resonators. The extremely important function of the varactors in the FET sources is to accurately control the amount of negative resistance introduced by the FET's. Clearly, too little negative resistance would compromise the filter performance, but on the other hand too much negative resistance could lead to stability problems. In the case of an unloaded resonator it can easily be shown that if the inductor has a series resistance of R_{ind} then a negative resistance $-R_{\text{ind}}$ is required. Hence one might expect that

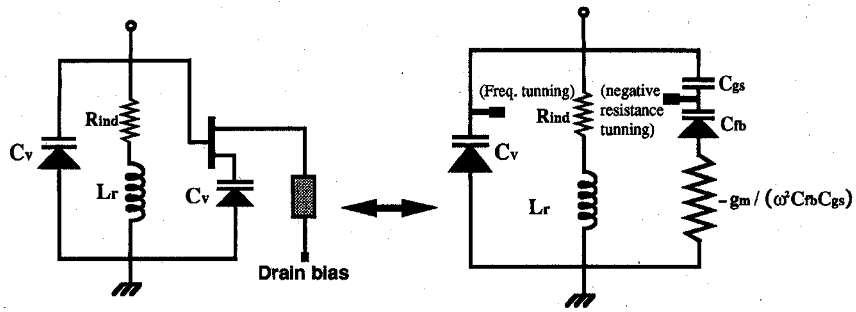
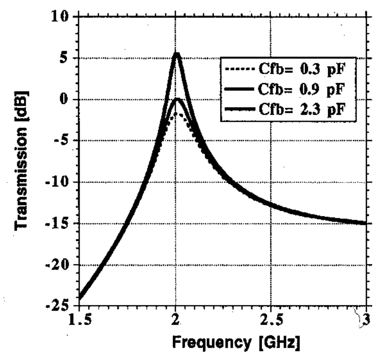
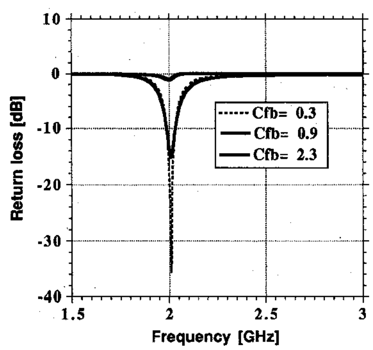


Fig. 5. Simplified equivalent circuit of a single active resonator.

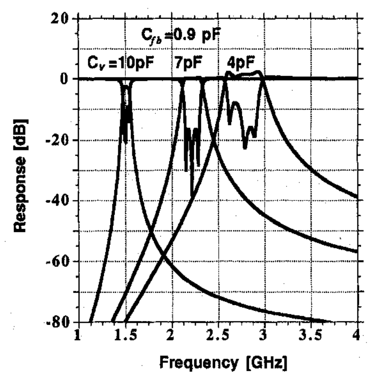


(a)

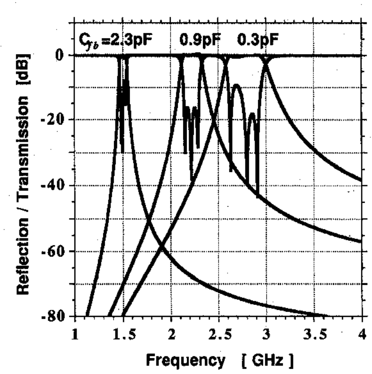


(b)

Fig. 6. Response of a loaded resonator as the source feedback capacitor is tuned. (a) Transmission, (b) reflection.



(a)

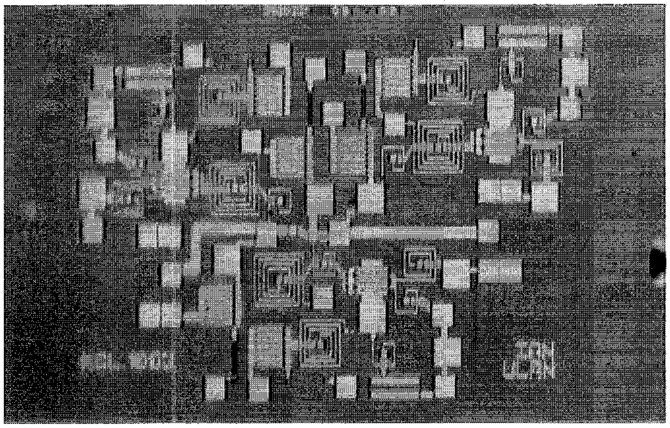


(b)

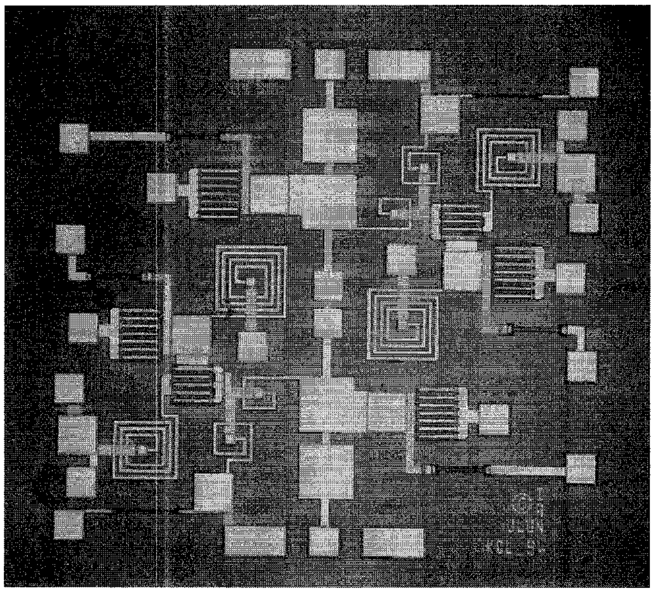
Fig. 7. (a) Tuning characteristics of varactor tuned filter when the negative resistance is fixed ($C_{fb} = 0.9$ pF). (b) Filter tuning characteristics when the negative resistance is varied (varactor tuned).

since a typical MMIC spiral inductor has say 4 Ω series resistance then the FET needs to provide only a fixed −4 Ω. In practice, the required value is larger and frequency dependant: The negative resistance circuit itself generates between −80 and −100 Ω throughout the passband. The flatness of the filter response depends on how the value of the negative resistance varies with frequency within the passband. In Fig. 6 it is shown how the feedback capacitance (the varactor diode) can be used to tune the response of a single coupled resonator. In this case the optimum value is 0.9 pF for a lossless resonator, and insertion gain can even be achieved.

To summarize these points, the common-source FET negative resistance circuits can use varactor tuning to control the negative resistance precisely. Since the common-base and common-gate variants require an inductor as the feedback element, there is less scope for electronic tuning. Incorporating a varactor diode for control of the negative resistance would create a series resonance which could lead to oscillation. The only available control is the bias point of the transistor itself, and varying this would lead to undesirable other changes in the resonator. The need for accurate control of the negative resistance is demonstrated in Fig. 7(a), in which a filter of the type shown in Fig. 4(b) has been simulated and tuned from 1.5 to 2.8 GHz with a fixed negative resistance. It can be seen that as the response is tuned down in frequency the negative resistance is no longer high enough, leading to high insertion loss, whereas when the filter is tuned up in frequency the negative resistance is too large, leading to a reflection coefficient greater than unity at some frequencies. In Fig. 7(b) the same simulations are carried out with the tuning on the



(a)



(b)

Fig. 8. Photographs of the fabricated MMIC filters. (a) 2.3 GHz three stage filter, (b) 4.7 GHz two-stage filter.

source varactors adjusted correctly. The insertion loss is now constant, the passband flatness is preserved and there is no danger of instability.

IV. Monolithic Filter Designs

Two and three resonator filters were designed to demonstrate the new technique experimentally. The GEC Marconi Materials Technology Ltd. (Caswell) F20 Foundry was used to fabricate the circuits. This process offers 0.5 μm gate-length ion-implanted MESFET's and through-substrate via-holes. For the three-stage filter the MESFET's are in a self-biased configuration for simplicity (to reduce the number of the bias lines) by employing a resistance in the source and by DC grounding the gate through an inductor which is also used as the inductive part of the active resonator. However, this self-biased configuration does not facilitate maximum tuning, so the two-stage design employs conventional biasing with separate gate bias lines. The drain bias circuit was optimized in order for the FET to give the required negative

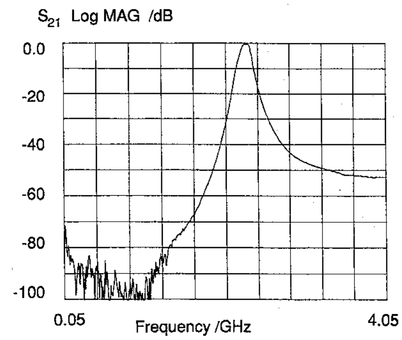


Fig. 9. Measured 2.3 GHz filter response.

resistance. The abrupt junction varactor diodes are realized using MESFET's with the source and drain fingers connected together [23], [24]. While these are not high Q varactors, the active resonator can easily compensate for their loss by introducing more negative resistance. However, their limited capacitance tuning ratio means that they serve mainly to optimize the filter characteristics and combat CAD inaccuracy and fabrication tolerances in order to achieve a successful first pass design, rather than to provide a highly agile frequency tuning performance. The monolithic two bandpass filters have been realized on a single 3 × 2 mm^2 chip for the three-stage and 2 × 2 mm^2 chip for two-stage and photograph of the fabricated chips are shown in Fig. 8(a) and (b), respectively.

V. Measurements of the MMIC Filter

The filter chips were measured using an HP8510$^{\text{TM}}$ automatic network analyser and Cascade Microtech$^{\text{TM}}$ wafer prober. In Fig. 9 the measured response of the three-stage filter shows it operating at 2.3 GHz center frequency with 120 MHz 3 dB bandwidth, with 0 dB insertion loss and less than ±0.1 dB ripple in the passband. The out-of-band rejection is as high as 100 dB at low frequencies and is over 50 dB up to 6 GHz. Furthermore, even up to 20 GHz there are no significant spurious transmission bands. The circuit was operated with a drain voltage of 3 V and drain current of 50 mA, giving a total DC power consumption of 150 mW.

Fig. 10(a), (b), and (c) show the measured response of the two-stage filter, which operates at 4.7 GHz center frequency with 400 MHz 3 dB bandwidth, tuneable insertion gain and less than ±0.1 dB ripple in the passband. The out-of-band rejection is as high as 70 dB at low frequencies and is over 23 dB up to 18 GHz. The circuit was operated with a drain voltage of 3 V and drain currents of 25 mA, giving a total DC power consumption of 150 mW. Two-tone third-order intermodulation characteristics have been measured using two sweepers and a spectrum analyzer: Fig. 11 shows the output spectrum from the filter for two −10 dBm tones. Saturation occurs at an input power of 0 dBm per tone, and the third-order intercept point is +11 dBm at the input. The filter could be designed to operate at higher power levels if back-to-back tuning varactors were employed [15]. The noise figure, measured with an HP8970 noise figure meter set up, is plotted in Fig. 12. This is one of the first reported noise figure results for an active filter employing negative resistance techniques,

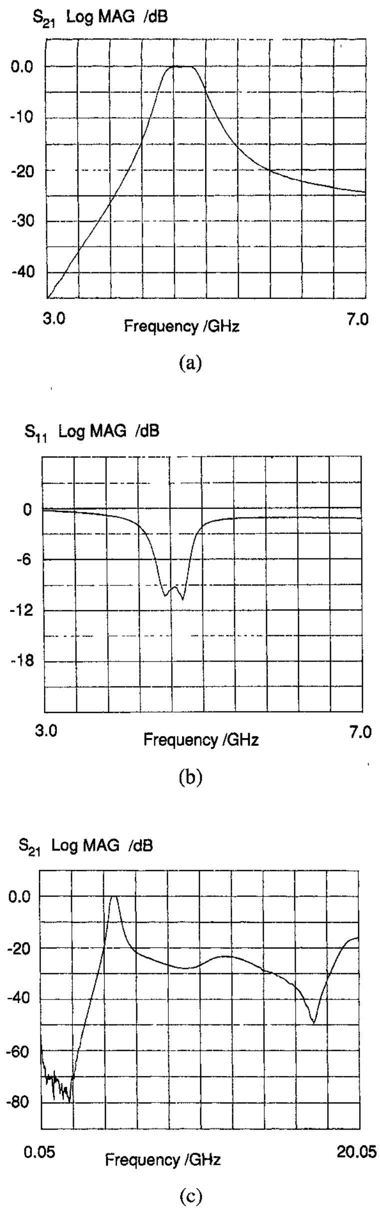


Fig. 10. Measured 4.7 GHz filter response. (a) Transmission, (b) return loss, (c) broad-band transmission response.

and shows that the noise figure is respectable, although as expected it is rather higher than for an amplifier. The stability of the filter over temperature has not been measured for this particular filter, but previous data for a filter using the same negative resistance circuit technique has shown that it is not a major limitation [25].

VI. Conclusion

The use of negative resistance transistor circuits to realize microwave active filters has been described, and the use of bipolar and FET devices has been compared. It has been shown that the FET can be used to realize a novel lossless active resonator which can be controlled easily and precisely using varactor diodes. This technique is ideal for MMIC applications, and two monolithic bandpass filters employing this technique have been demonstrated. The filters have center frequencies of 2.3 GHz and 4.7 GHz with a 120 MHz and 400 MHz 3 dB bandwidth, respectively. Both have a measured insertion loss of 0 dB. The 4.7 GHz design has been fully characterized in terms of linearity and noise performance. It is expected that this active filter technique can contribute significantly to the further integration and miniaturisation of advanced microwave subsystems.

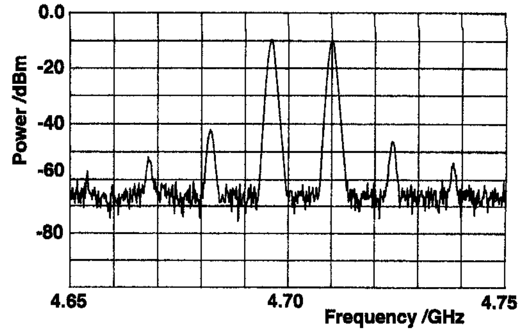


Fig. 11. Output spectrum from the 4.7 GHz filter for two −10 dBm input tones.

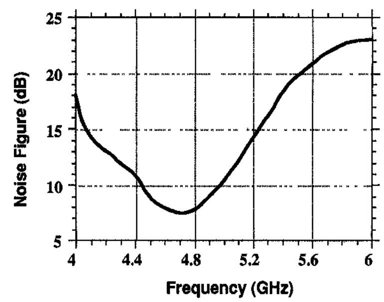


Fig. 12. Measured noise figure of the 4.7 GHz filter.

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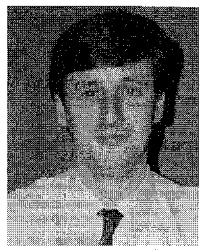

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From 1984 to 1986 he was employed at Plessey Research (Caswell) in the MMIC Research Group, where he worked on MMIC mixers, on-wafer measurement techniques, and FET characterization. In 1986 he returned to King's College, first as a Research Assistant, working on the T-SAT mobile communications payload. He is currently a Reader in Microwave Engineering at King's College and leader of the MMIC Research Team in the Communications Research Group. His research interests encompass all aspects of the design and application of MMIC's. He recently edited *MMIC Design* published by the Institution of Electrical Engineers.