AIP Journal of Applied Physics

HighQ magnetostatic surface wave planar yttrium iron garnet resonator

D. K. De

Citation: Journal of Applied Physics **64**, 2144 (1988); doi: 10.1063/1.341726
View online: http://dx.doi.org/10.1063/1.341726
View Table of Contents: http://scitation.aip.org/content/aip/journal/jap/64/4?ver=pdfcov
Published by the AIP Publishing

High-Q magnetostatic surface wave planar yttrium iron garnet resonator

D. K. De[a)]

Department of Electrical Engineering, The University of Texas at Arlington, Arlington, Texas 76019

(Received 28 September 1987; accepted for publication 25 April 1988)

Experimental results on the magnetostatic surface wave resonance characteristics of planar yttrium iron garnet resonators are given. Single-mode resonance with wide tunability range and high Q have been achieved. Q as high as 3500 and insertion *loss* as low as 13 dB with off-resonance rejection as high as 15 dB have been achieved at 6 GHz. A microwave oscillator built with the resonators and a laboratory-made hybrid amplifier provided oscillations in the range 3–5.3 GHz with phase noise of -105 dBc/Hz at 10 KHz offset *from* the carrier frequency.

INTRODUCTION

Magnetostatic wave propagation (MSW) in EPI yttrium iron garnet (YIG) film satisfies the equations: $\nabla \times h = 0$ or $h = \nabla \psi$ and $\nabla \cdot e = 0$ where h and e are the small signal magnetic field and electric field vectors. The wave can propagate in three modes such as surface, forward volume wave, and backward volume wave.[1] The surface wave mode (MSSW) is excited when the bias field lies on the surface of the YIG plane such that the field direction is normal to the propagation direction. For backward-wave volume wave (MSBVW) the field is in the direction of the wave propagation, and for forward volume wave (MSVW) the field is perpendicular to the YIG surface.

Magnetostatic wave (MSW) devices form the basis of a new emergence of microwave analog signal processing. Among these devices, MSW resonators and the corresponding loop transducers can be fabricated using conventional UV photolithography. It provides broad frequency tunability (2–18 GHz) by an external magnetic field. The insertion loss is low, and the Q is moderate, whereas for acoustic wave resonators one needs submicron photolithography; though Q is usually high the insertion loss is also very high at microwave frequencies. It is well known that YIG sphere resonators operate well at microwave frequencies, but require tedius and expensive fabrication procedures. MSW techniques, however, provide a simpler means of obtaining very-high-Q resonators for MSW oscillator and complex filter functions.

Magnetostatic surface waves are predominantly magnetic, dispersive electromagnetic waves[1] that propagate with the field maxima at the ferrite top and bottom surfaces. With a conducting plane close to one of the surfaces the wave dispersion for the waves propagating through the top and bottom surfaces are different. The forward MSSW mode is one for which the wave energy is confined to the top surface of the YIG and the reverse mode is one for which the energy is confined to the bottom surface, and this has a higher insertion loss than the forward mode. It is the reverse mode of MSSW that gives rise to the MSSW resonances.

During the early investigations of MSW it was found that resonances occur in rectangular slabs of YIG.[2] MSW resonators, however, have been built using thin films of YIG with grating reflectors.[3–6] The grating elements which act as reflectors to MSW waves are formed by several types of discontinuities, such as (i) chemically or ion-beam-milled etched groove arrays, or (ii) metal dot arrays. Resonators other than grating elements have been investigated by utilizing the straight edges produced by cutting the YIG films with a wafer saw.[7,8] Though high-Q resonances have been obtained, the insertion losses are higher and the resonances are multimoded, and mode crossing occurs for tunability over wide frequency ranges.

The present investigation on a grating resonator shows that such a resonator can be made with the following characteristics:

(1) Wide tunability range.

(2) Very-high-Q and high off-resonance rejection.

(3) Only single-mode resonance which is essential for the use of a resonator in a microwave oscillator.

(4) Insertion loss lower and Q higher than that reported in the literature.[2–6]

DESCRIPTION OF THE RESONATOR

A schematic of a two-port MSSW resonator along with its equivalent circuit is shown in Fig. 1. The two-port planar YIG (the YIG film is grown epitaxially on 250-μm GGG substrate) resonator consists of a pair of distributed grating elements [in the form of two etched groove arrays consisting of land (L) and groove region (G)] which are separated by a propagating space (S) that forms a cavity. The input and output microstrip loop transducers which are fabricated on a quartz substrate (Q) are usually placed within the space (S) such that the structure (A) is on top of the structure (B) in flipped configuration. The resonators are tuned over different frequencies by adjusting the bias magnetic field shown as H_{dc} in Fig. 1. In general, H_{dc} versus frequency relation is approximately linear.[7] The equivalent circuit of a typical MSSW resonator is shown in Fig. 1(c). The corresponding parameters can be obtained from experiments and using the following equations:

$$\text{Insertion loss (dB)} = -20 \log r, \tag{1}$$

$$R_1 = R_0(1-r)/(1+r), \tag{2}$$

[a)] Present address: Radiology Department, The Bowman Gray School of Medicine, 300 S. Hawthorne Road, Winston Salem, NC 27103.

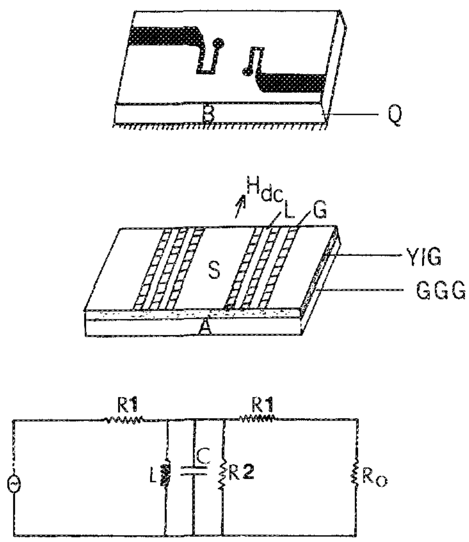


FIG. 1. MSSW resonator structure with its equivalent circuit (c). The dimension of a typical resonator is 25 mm × 1 mm × 264 μm (14-μm YIG film on 250-μm GGG substrate).

$$R_2 = 2R_0r/(1-r), \quad (3)$$

$$R_0 = 50\ \Omega, \quad (4)$$

$$Q = 2R_0C\omega/(1+r)^2, \quad (5)$$

$$L = 1/\omega^2 C. \quad (6)$$

In order to design a resonator, it is essential to know the dependence of quality factor (Q), insertion loss (IL), and off-resonance rejection (R), etc., on the resonator physical parameters like (1) YIG film thickness h, (2) number of etched grooves N, (3) etch depth d, (4) separation between the two arrays a. Initial theoretical investigations on these dependencies based on the model of cascaded transmission line sections due to each land and groove region has been done by Snapka.[6] In the present experimental investigation, a systematic variation up to $\pm 20\%$ of the theoretical values[6] of h, d, a, and N were allowed in order to optimize Q, IL, and R. A large number of resonators were fabricated and tested with different loop transducer configurations. The resonator widths were successively reduced from 3 to 1 mm.

EXPERIMENTAL RESULTS

It was found that although very high Q (3000 or more at 6 GHz) can be obtained with 1-in-long by 3-mm-wide resonator, IL was too high (34 dB) and R was too low (3–6 dB). Moreover, clean tunability was limited to only 5 GHz. Improvement in IL was made by reducing the resonator width and consequently the length of the transducer microstrips. By matching, IL can be reduced substantially and R is also increased.

This present investigation shows that it is possible to meet all the four criteria mentioned in the Introduction with 1-mm-wide and 25-mm-long resonators built with EPI YIG films on 250-μm-thick GGG substrate (YIG thickness $h \sim 14\ \mu$m, etch depth $d/h = \sim 2.5\%$, number of bars = 65, array separation = 2.4 mm). The width of the land region is determined by $\lambda/4$, λ being the MSSW wavelength of operation. The groove width is given by: $g = \lambda(1 - d/h)/4$. The loop spacing of transducer microstrips is $\lambda/2$. In this work, λ was chosen to be 300 μm. The spacing between the arrays was kept an integral multiple of λ for different resonators studied in this work.

The measured loaded Q was 800–3000 over 3–6 GHz for a 3-mm-wide resonator. Reduction in the width of the resonator to 1 mm reduced the Q to the range 550–2000 over 3–6 GHz, still much higher than the YIG sphere resonator. The insertion loss with matched transducer was in the range 6–14 dB (Figs. 2–4). The off-resonance rejection ~10–15 dB, higher at higher frequencies. The 1-dB power compression points at 3 and 6 GHz were − 10 and 4 dBm. The magnetic field tunability of the resonator in the range 3–6 GHz is shown in Fig. 4. These resonators are usable at much higher frequencies, however, with higher Q and insertion loss as well. The resonance throughout the entire frequency range is single moded, and consequently no mode crossing is observed. Mode crossings were found to occur even in the case of the best straight-edge MSSW resonators.[7] It may be mentioned that 3-mm-wide etched grooved resonators did not have clean tunability beyond 5.0 GHz. This is because of the appearance of the uniform resonance mode higher than the MSSW resonance (at the low-frequency side of the MSSW resonance).

MATCHING THE LOOP TRANSDUCER

As mentioned earlier, initially 3-mm-wide by 1-in-long resonators were investigated along with 3-mm-long microstrip loop transducers. Though the resulting Q were 800–3000 in the frequency range 800–3000, the corresponding insertion loss was very high 15–34 dB. Using 1-mm-wide resonators, the insertion loss was found to be reduced to 8–20 dB before matching. In order to reduce the insertion loss further, a 1-mm-long microstrip transducer was electrically matched to 50-Ω input-output load, at resonance. For matching, S_{11} parameters (both magnitude and phase) were determined at resonance over 3–6 GHz at a 200-MHz interval, with respect to corresponding 1-mm transducer electrical short obtained by grounding at the input of the microstrips. A Hewlett-Packard network analyzer was used for this purpose.

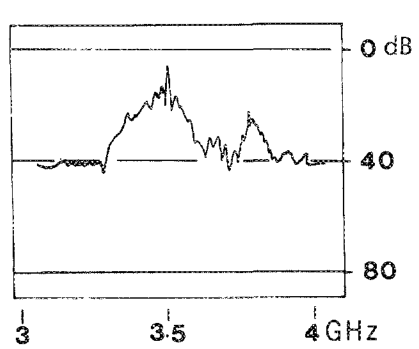


FIG. 2. 6-dB resonance insertion loss at $f = 3.5$ GHz; 1.5-mm-wide resonator $K_{2'}$ used with 1-mm matched transducer. The high-frequency-side resonance, however, is not observed for other resonators of 1 mm width.

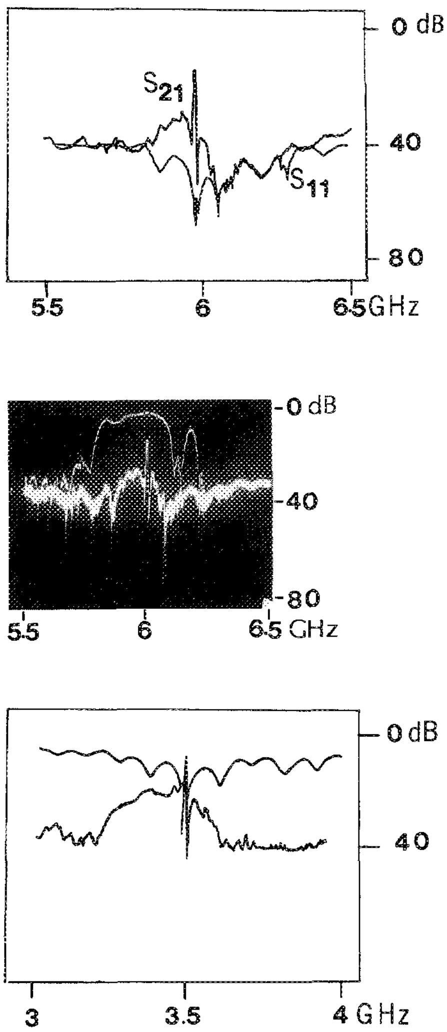


FIG. 3. (a) Resonator response at 6 GHz. 1-mm resonator K_2 is used with 1-mm matched transducer. It is seen that the S_{11} notch (upper curve) coincides with the S_{21} peak (bottom) trace. Sweep range = 1 GHz. (b) Resonator response at 6 GHz from 1.5-mm-wide resonator $K_{2'}$. $K_{2'}$ is obtained from the same batch of K_2 resonators after increasing the etch depth from 2.5% to 5.5%. (c) The response at 3.5 GHz from a resonator with $h = 16$ μm, $d = 0.7\,\mu$m, $S = 3.6\,\mu$m with a matched transducer. The S_{21} peak coincides with the S_{11} peak as is expected for good matching. Before matching the latter was expected for good matching. Before matching the latter was occurring at 83 MHz higher. Matching improved the insertion loss in this case by 7 dB.

S_{11} parameters were plotted on a Smith chart (Fig. 5) and primary matching design principles were followed as provided by Hewlett-Packard application note 154. As matching over an octave band 3–6 GHz is difficult, we tried finally to optimize the matching through COMPACT (computer optimization of microwave passive and active circuits). The circuit which optimized the matching is shown in Fig. 6 (built on 250-μm quartz substrates, $\epsilon_r = 3.8$). The optimized S_{11} parameters within 4000–5850 MHz are also shown in chart A. We were not very concerned about optimization in the range 3–4 GHz, because the insertion loss at resonance was within 6–12 dB even with an unmatched 1-mm transducer. Following standard equations[9] for microstrip circuits and the equivalent circuit of the COMPACT optimized (Fig. 6), the length l_L and the width w of the inductor microstrip were calculated to be 1.1 mm and 131 μm. The matched transducer, with 50-μm loop microstrips was then fabricated through the usual UV photolithographic technique. The output transducer was a mirror image of the input transducer shown in Fig. 6. After matching, the following improvements have been observed. The equivalent circuit of the resonator transducer combination is also shown in Fig. 1.

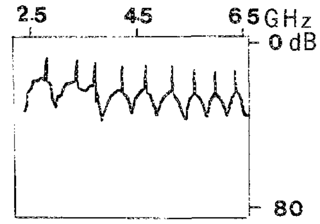


FIG. 4. Tunability of resonator response over 2.5–6.5 GHz with matched transducer after two months of continuous use.

RESULTS OF TRANSDUCER MATCHING

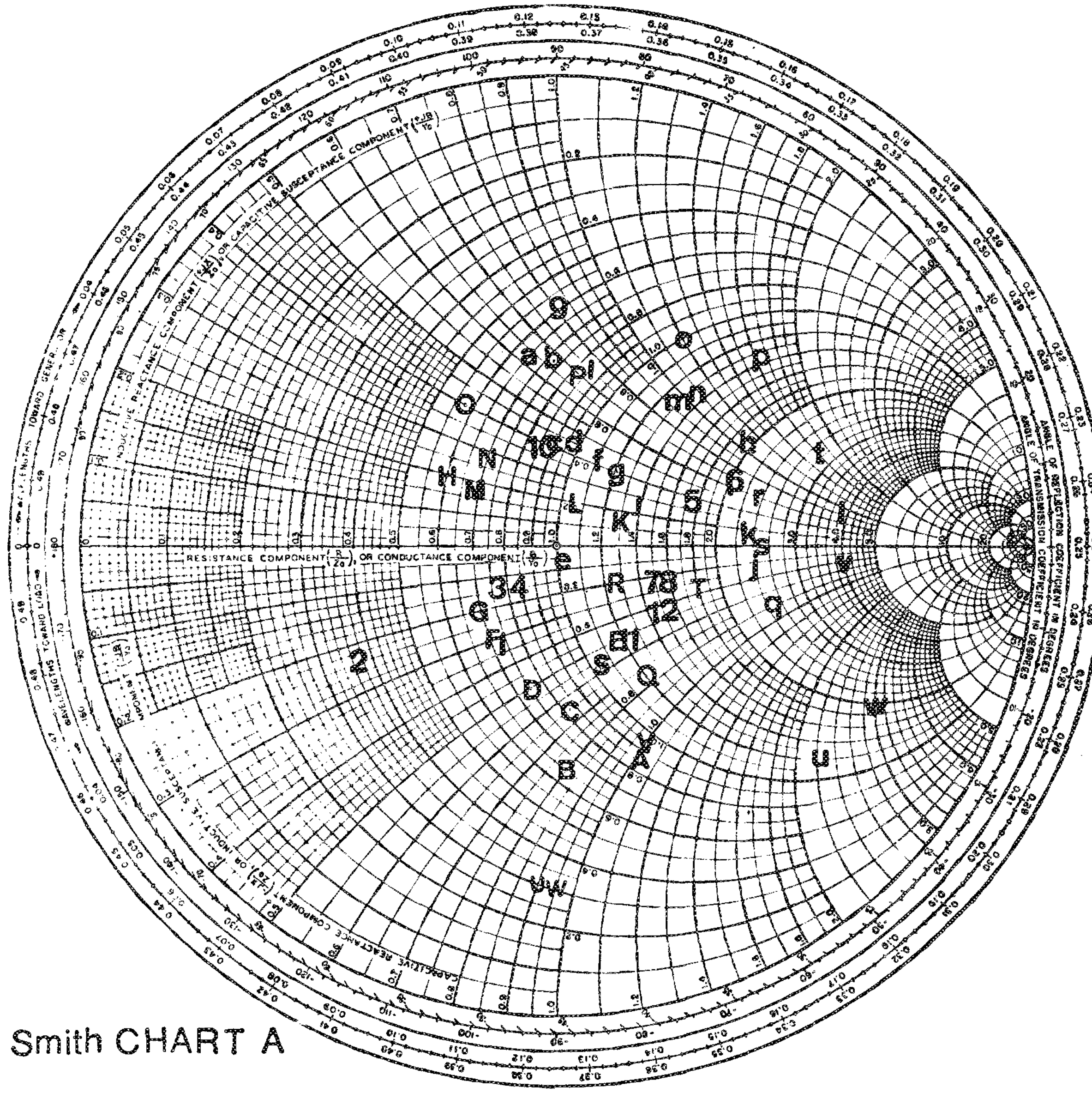
(1) It has been possible, in general, to reduce the insertion loss by 6 dB or more (in some special cases by 7–8 dB), i.e., the insertion loss at resonance with most of the 1-mm resonators can be kept at 8–14 dB over 3–6 GHz. At 3 GHz, insertion loss as low as 6 dB at resonance has been obtained for some resonators (Fig. 2). Relative improvements at matching can be seen in Figs. 3(a) and 3(c) by noting that the S_{21} (resonance) peak coincides with S_{11} (reflection peak), which does not occur in the case of unmatched resonator-transducer combination. With an unmatched resonator the S_{21} peak does not coincide with the S_{11} peak, e.g., in Fig. 3(c) (for a different resonator), before matching the S_{11} peak was at 83 MHz higher than the S_{21} peak and the corresponding insertion loss was about 8 dB higher.

(2) The off-resonance rejection has been improved by 4 dB, i.e., 10–14 dB within 3–6 GHz (Fig. 4). Before matching it was around 7–10 dB.

(3) The matched S_{11} parameters have been found to be quite low (0.150, around 5–6 GHz, compared to those of the computer values (Fig. 5). In the 3–5-GHz range the measured values were closed to the optimized values.

(4) Tunability also improved after matching. With most of the resonators, matched transducers did not provide higher-Q resonance.

In this work, single modedness of the resonance over 3–6 GHz was determined by finding out the phase shift at the resonance as the resonator is tuned over 3–6 GHz with respect to the phase at 3 GHz. For this the phase shift due to the transmission line of the transducers was taken into account. The only resonance phase shift of 10°–12° could be recorded over 3–6 GHz. Finally, it may be mentioned that one of the resonators K_2 had been used (after nearly four months of fabrication of the resonator and the transducer) in a feedback loop with a laboratory-made hybrid microwave amplifier for oscillation in the range 3–5.3 GHz. The amplifier had a gain of 18 dB in that range, with gain flatness of 3 dB and noise of 2 dB level. The microwave oscillation produced, had a phase noise of − 105 dBc/Hz at 10 KHz offset from the carrier.[10]



Smith CHART A

FIG. 5. Measured S_{11} parameters for the resonator I (7-μm YIG film) with 1-mm unmatched transducer are shown in alphabetic orders denoted by small letters at frequencies: 3.0, 3.2, 3.4, 3.6, 3.8, 4.0, 4.2, 4.4, 4.6, 4.8, 5.0, 5.1, 5.2, 5.3, 5.4, 5.5, 5.55, 5.65, 5.75, 5.85, 5.95, and 6.0 GHz. Capital letters show measured S_{11} at above frequencies with matched transducer. Numbers show S_{11} measured with another resonator (K_2) built with 14-μm YIG film, with a matched transducer in the range 3–6 GHz at steps of 0.2 GHz.

DISCUSSION

Earlier works[2–5] used high d/h (as high as 6.4% to 11%) to obtain high Q. Our works show that d/h basically helps to obtain a single-moded resonance with large off-resonance rejection but does not improve Q beyond certain d/h, e.g., Q was found to drop from 2000 to 1300 at 6 GHz when a resonator from the batch K_2 was etched further to increase d/h from 2.5% [Fig. 3(a)] to 5.5% [$K_{2'}$ in Fig. 3(b)]. The corresponding insertion loss (14 dB) only slightly improved by ~1 dB and the off-resonance rejection dropped from 14 to 12 dB [Fig. 3(b)]. For a given resonator transducer combination, Q at the resonance also depends on the position of the transducer relative to the spacing (S) between the two arrays of bars and grooves. By reducing the separation between the two arrays, the spurious response or side lobes can be diminished to a great extent; however, the loaded Q is also theoretically predicted to be reduced.[5] Reduction in Q may be compensated by properly choosing the etch depth ($d/h \sim 2$–3%), land groove width as well as the position of the transducer with respect to the cavity spacings. Our works show that small array separation (<2 mm) does not provide high off-resonance rejection. Higher off-resonance rejection may be possible by carefully choosing the separation of the loop strips. Experimentally, further improvements may be attempted by giving plus or minus small varia-

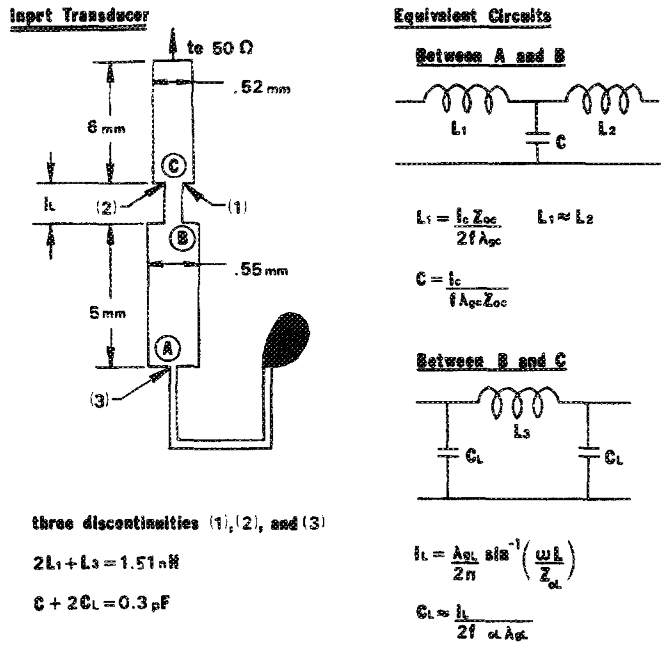


FIG. 6. Microstrip equivalent circuit obtained after matching through COMPACT. The dimension of the loop microstrip of the transducer is 1 mm × 50 μm. This is designed for 300-μm MSSW wavelength. Input and output transducers are separated by 2.4 mm. They are placed within the cavity of the resonator.

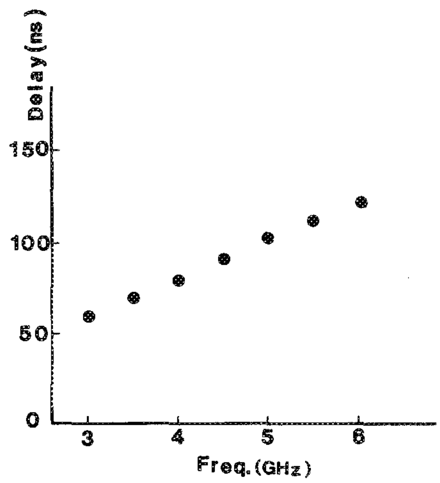


FIG. 8. Time-delay measurements at resonance in the range 3–6 GHz. Resonator 1 mm wide (with 7-μm YIG film) and 1-mm matched loop transducer are used for measurements.

tions about the length and width of the inductor microstrip (Fig. 6) to reduce the S_{11} at high frequencies. Because of metal loading by the YIG, the effective wavelength λ_{eff} of the wave launched by the transducer differs from that expected on the basis of the equation $\lambda_0 = 2s$, where s is the separation between the two microstrips of a loop. If this separation is chosen carefully, the insertion loss will also improve further and it will be less than 14 dB at 6 GHz. This may be true for a 100-μm wavelength short resonator (the advantage of using small MSSW wavelength is that the resonator length can be shortened to ~ 8 mm, keeping the number of bars in each array ~ 60, and Q is very high for shorter wavelengths because of increased delay time τ) as can be seen from Fig. 7, where the resonance has been placed out of the top of the transducer passband. Otherwise, the insertion loss could have been 16 dB at 6 GHz even with an unmatched transducer. This problem together with the effect of the surface quality of the YIG film on Q and IL, etc.; remains to be further investigated. It may be mentioned that at higher bias field (which corresponds to higher resonance frequency) the transducer passband shrinks because of the narrowing down of the difference between the upper and lower cutoff frequencies in a ferrite[1] for the MSSW wave. Because of this, the matching of wavelength launched by the transducer with the periodicity of the resonator grating becomes very important. A small mismatch will place the resonance peak at a different position other than the top of the passband, and this will add to the insertion loss significantly at high frequency. Change of wavelength because of the metal loading by the YIG also needs to be taken into account. These above-mentioned problems are found to be very important for short resonators [6–8 mm long and 1 mm wide, 14-μm YIG film on 250-μm GGG substrate operating at low MSSW wavelength (100 μm)]. Such resonators were found to have Q in excess of 3500 at 6 GHz (Fig. 7). If the above-mentioned corrections could be made then such short resonators can have very low insertion loss and high off-resonance rejection and would be highly useful in microwave oscillator.

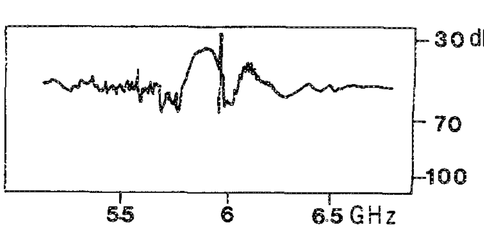


FIG. 7. Narrow resonator (7 mm × 1 mm × 263 μm) designed for 100-μm wavelength. The loop transducer was also designed for the same wavelength. However, because of metal loading by the YIG the corresponding wavelength is changed and the resonance peak is shifted out of the top of the transducer passband. The insertion loss with unmatched transducer is 26 dB and the Q is 3500.

It may be mentioned that we have carried out time-delay measurements at resonance by finding out the phase shift $\Delta\Phi$ when the frequency is changed by $\Delta f\,(=5\text{ MHz})$ about the resonance point. The delay τ is measured through the equation

$$\tau = \left(\frac{1}{2\pi L}\right)\frac{\Delta\Phi}{\Delta f},$$

where L is the spacing between the input and output transducers. The measured time delay over the frequency range 3–6 GHz is shown in Fig. 8 for the resonator I and a matched 1-mm microstrip transducer. The resonator tranducer combination was set for $Q = 500$–1700 in the range 3–6 GHz with corresponding insertion loss 8–16 dB. Increase of insertion loss by 2 dB (higher than 14 dB obtained from fresh transducer) is possibly due to the deterioration of the transducer microstrip and the OMS connectors upon use over a span of three months. It is found that the delay varies linearly (Fig. 8) with resonance frequency in the range 3–6 GHz. It should be noted that different delay times correspond to different bias fields at the corresponding resonance frequencies. The delay τ of a resonator is related to its Q by $Q = \pi\tau f$. Though the measured Q vs f of a resonator in most cases have been found to be approximately linear, the loaded Q is about 25% smaller than that calculated from $Q = \pi\tau f$. This may be associated with the propagation loss in the YIG film and its loading by the metal transducers.

Finally it may be mentioned that one can obtain the equivalent circuit parameters [Fig. 1(c)] of the resonator K_2 from the Eqs. (1)–(6), using loaded Q of 1600 and IL = 14 dB at 6 GHz, as $R_1 = 33\ \Omega$, $R_2 = 25\ \Omega$, $L = 0.22$ pH, and $C = 3.2$ nF.

CONCLUSIONS

(1) By reducing the width of the resonator and the length of loop strip of the transducer, the insertion loss at resonance is reduced significantly. This is also reflected in the lower S_{11} parameters (see the Smith chart plot A of S_{11}).

(2) By matching the transducer the insertion loss has been reduced further by 6 dB. The off-resonance rejection has improved by at least 4 dB; tunability has improved. Q, however, did not improve.

(3) Q of resonators (as studied in this work) depends mainly on (i) exact reproduction of the mask on YIG (i.e., uniform land groove spacings throughout the arrays) as well as the transducer on quartz substrate; (ii) a large number of bars per array, typically ~ 60; (iii) $d/h = \sim 2.5\%$.

(4) High-resonance Q and low insertion loss are possible with MSSW resonators.

ACKNOWLEDGMENTS

The author thankfully acknowledges financial support and guidance given by Professor J. M. Owens. The author is also grateful to Professor R. L. Carter for encouragement and helpful suggestions during the progress of this work.

[1] Microwave Propagation in Ferrimagnetics, M. S. Sodha and N. C. Srivastava (Plenum, New York, 1981), p. 170.

[2] L. K. Brundle, Electron. Lett. 4, 132 (1968).

[3] W. R. Brinlee, J. M. Owens, C. V. Smith, Jr., and R. L. Carter, J. Appl. Phys. 52, 2276 (1981).

[4] J. P. Castera and P. Hartmann, in Proceedings of the 1981 RADC Microwave Magnetics Technology Workshop, RADC TR 83-15 (1983), pp. 218–229.

[5] J. H. Collins, J. D. Adam, and Z. M. Bardai, Proc. IEEE Lett. 65, 1090 (1977).

[6] E. P. Snapka, M. S. thesis, The University of Texas at Arlington, Arlington, Texas, 1978.

[7] K. W. Reed, A. K. Reddy, and W. A. Davis, Int. Microwave Symp. Dig. IEEE MTT, 519 (1985).

[8] E. Huijer and W. Ishak, IEEE Trans. Magn. MAG-20, 1232 (1984).

[9] T. C. Edwards, *Foundations for Microstrip Circuit Design* (Wiley, New York, 1981), p. 216.

[10] R. L. Carter, J. M. Owens, and D. K. De, IEEE Trans. Microwave Theory Tech. MTT-32, 1671 (1984).