

Miniature SAW Antenna Duplexer for 800-MHz Portable Telephone Used in Cellular Radio Systems

MITSUTAKA HIKITA, SENIOR MEMBER, IEEE, YOSHIKATSU ISHIDA,
TOYOJI TABUCHI, AND KAZUHITO KUROSAWA

Abstract—A new low-loss, high-power SAW filter has been developed for an antenna duplexer used in an 800-MHz portable telephone transceiver. The new ladder-type configuration uses SAW resonators and capacitors between electrode fingers and the earth. It is fabricated on a 2-mm-square LiTaO₃ chip. This configuration provides insertion loss as low as 1.0 ~ 1.2 dB and output power up to 2 W at 800 MHz. Experimental results of a miniature antenna duplexer module using this filter as a transmitter filter and the previously published high-performance SAW filters as receiver filters are also presented. This module also includes a receiver low-noise amplifier, yet its size is under 8 ml.

I. INTRODUCTION

A PORTABLE telephone (cellular radio) requires high-performance, miniature devices. SAW technologies are noteworthy due to their small device size compared with conventional technologies [1], [2].

An antenna duplexer consists of transmitter and receiver filters connected in parallel via appropriate transmission lines, as shown in Fig. 1. The transmitter filter must have low-loss, high-power characteristics. The receiver filters must provide low loss and high sidelobe suppression. In a conventional duplexer, each filter is constructed with 4 ~ 7 cascade-connected semicoaxial resonators containing high-dielectric ceramics. Thus the duplexer is one of the largest circuit components in a two-way transceiver, such as a portable telephone.

SAW filters are miniaturized filters given that the SAW wavelength is about 10^{-5} times shorter than that of electromagnetic waves [3], [4]. Interdigital transducers (IDT's), which are constructed with 1.1 ~ 1.2- μm -wide electrode fingers at 800 MHz, are easily fabricated using standard photolithographic techniques, such as those used with silicon IC's [1]. However, conventional SAW filters have rather large insertion losses of 10 ~ 20 dB. Thus they cannot withstand input power above about 15 dBm at 800 MHz.

This paper examines a new low-loss, high-power transmitter SAW filter at 800 MHz from initial specifications to final device operation. It describes a new filter configura-

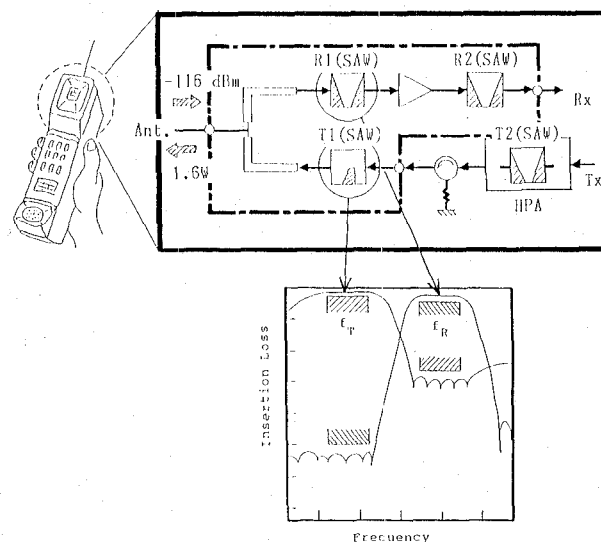


Fig. 1. Block diagram for SAW antenna duplexer.

tion which provides insertion loss as low as 1.0 ~ 1.2 dB and high-power characteristics of up to 2 W. Computer simulation procedures and experimental results are also presented.

The authors have already reported several new technologies for achieving high-performance SAW filters [2]. These include a new phase-weighting approach, image-impedance connection of IDT's, and a new IDT repetition structure with a tapered number of finger pairs [2].

Receiver filters which provide low loss and high sidelobe suppression are achieved using combinations of these new technologies. Experimental results for a miniature SAW duplexer module using the newly developed high-power SAW filter as a transmitter filter and the above high-performance filters as receiver filters are also presented. The duplexer also includes a low-noise amplifier and a second receiver filter. However, its size is under 8 ml.

II. ANTENNA DUPLEXER

An antenna duplexer consists of two kinds of filters: transmitter and receiver filters. Two filters, T1 and R1, are connected in parallel via appropriate transmission lines, as Fig. 1 shows. Such a duplexer requires (1) low-loss (below 2 dB) and high-power characteristics for the transmitter

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M. Hikita and T. Tabuchi are with the Central Research Laboratory,
Hitachi, Ltd., Kokubunji, Tokyo, 185 Japan.

Y. Ishida and K. Kurosawa are with the Hitachi Tokai Works, Ouaza,
Katsuta, Ibaragi, 312 Japan.

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filter, and (2) low loss (below 4 dB) and high sidelobe suppression for the receiver filters.

A conventional duplexer is usually formed with T1 and R1. Each filter is constructed of semirigid coaxial resonators (4 ~ 7 resonators for each filter) containing high-dielectric ceramics. The other filters, T2 and R2, are also constructed of semirigid coaxial resonators. They are assembled in a motherboard together with other active and passive circuit elements. Thus the duplexer and the other filters have occupied the largest circuit areas in two-way transceivers, such as those in mobile and portable telephones.

The authors have developed two kinds of RF modules for portable telephones. One is a miniature SAW antenna duplexer module containing three SAW filters (T1, R1, and R2) and a low-noise amplifier. The other is MOSFET high-power amplifier module which also contains a SAW filter, T2. This filter is mounted on a miniature TO-18 hermetically sealed package. This paper treats the former theoretically and experimentally. The latter will be published in the near future.

III. NEW CONFIGURATION FOR LOW-LOSS, HIGH-POWER SAW FILTER

A. Power Limitation of Conventional SAW Filters

Conventional transversal SAW filters convert electromagnetic energy to SAW energy using the electrostatic field pattern of input IDT coupling to the SAW through the piezoelectric effect [3]. Thus maximum input power is limited to 10 ~ 15 dBm in 800 MHz conventional transversal SAW filters. This is because 1.1 ~ 1.2 μm electrode fingers are used in IDT's at 800 MHz. The mechanical migration caused by the large vibration amplitude of SAW's degrades these IDT electrode fingers.

Our basic experiments herein use aluminum electrode fingers doped with 1.5 ~ 2.5 percent copper. This ensures about 5 dB improvement for input power. This improvement was first pointed out by Latham *et al.* [5] using SAW resonators with groove reflectors, fabricated on a quartz substrate. Ebata *et al.* [6] have also experimented with SAW resonators with metal grating reflectors.

However, EIA standards define the radiation power from class II portable telephones as 1.6 W. Thus the duplexer must withstand ~ 35 dBm input power. To achieve over 15 dB improvement, a new filter configuration must be developed.

B. New Ladder-Type SAW Filter

The newly developed low-loss, high-power SAW filter is constructed with electrically cascade-connected SAW resonators. This is shown in Fig. 2. Simple outer matching circuits are also introduced to input and output terminals. This configuration's advantages are (1) low insertion loss, (2) high-power characteristics, and (3) sharp cutoff frequency responses at frequencies higher than the pass-band.

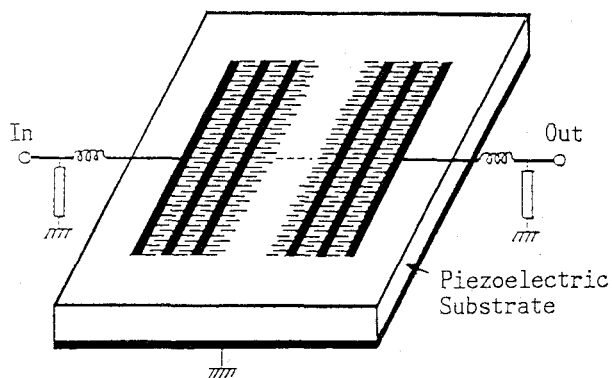


Fig. 2. Configuration for high-power SAW filter.

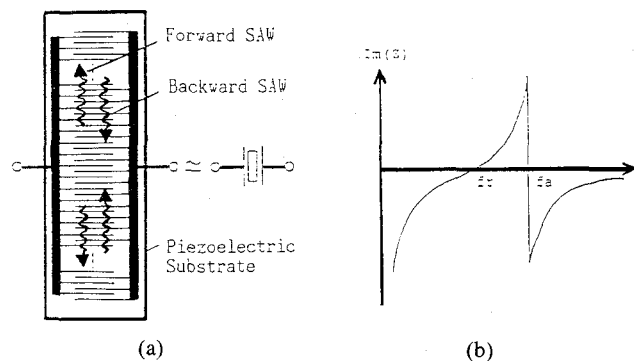


Fig. 3. Wide-band SAW resonator. (a) Unit resonator. (b) Frequency characteristics.

Each SAW resonator consists of an IDT with 300 ~ 400 electrode finger pairs, as Fig. 3 shows. Backward SAW's with almost the same amplitude as that of forward SAW's can exist within an IDT with this many finger pairs. This is because in a periodic structure, forward- and backward-traveling SAW's interact due to internal reflections.

This configuration differs from the well-known conventional SAW resonators, i.e., reflector-type resonators, as follows. The configuration in Fig. 3(a) provides a very wide frequency band resonator. This phenomenon is very similar to that of the distributed feedback laser (DFB). An example of this SAW resonator's impedance characteristics is shown in Fig. 3(b). The resonator's antiresonant/resonant frequency ratio (f_a/f_r) is much larger than that of conventional resonators.

The transmitter filter is designed to reduce the receiver band noise generated from the high-power amplifier (HPA). In general, the transmitter frequency band of cellular radio systems is allocated at the lower side of the receiver frequency band. Thus the ladder-type filter using SAW resonators and capacitors seems to suit the purpose. Here the antiresonant frequency of each resonator is assigned within the receiver band. This filter is shown in Fig. 4.

The configuration shown in Fig. 2 is a one-chip realization of this ladder-type filter. C_i , where $i = 0, 1, 2, \dots, n$, can be achieved by the sum of the capacitances between resonator electrode fingers and the earth and of the capacitances between resonator-resonator coupling pattern areas and the earth.

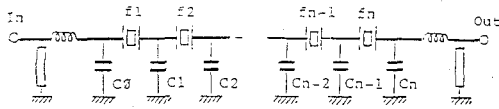


Fig. 4. Ladder-type equivalent circuit.

C. Frequency Characteristics for New Filter

To explain frequency responses for the new filter more precisely, simplified equivalent circuits are introduced for the configuration shown in Fig. 2. An approximate equivalent circuit at all frequency ranges is shown in Fig. 5(a).

At frequencies lower than the resonant frequencies of the resonators, each resonator's impedance becomes only capacitive due to the resonator's many electrode finger pairs. Thus an equivalent circuit of the filter is represented approximately by Fig. 5(b).

At frequencies near the resonant frequencies of the resonators, each resonator's impedance approaches a series circuit of inductance and capacitance due to the mechanical reaction of SAW vibrations. This produces the equivalent circuit shown in Fig. 5(c).

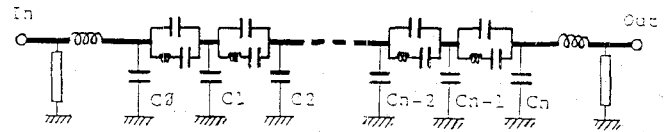
At frequencies near the antiresonant frequencies of the resonators, each resonator's impedance approaches a parallel circuit composed of the capacitance between electrode finger pairs and inductance due also to mechanical reaction of SAW vibrations. This produces the equivalent circuit shown in Fig. 5(d).

At frequencies sufficiently higher than the antiresonant frequencies of the resonators, each resonator's impedance is again represented only by the capacitance between electrode finger pairs, resulting in the simple equivalent circuit of Fig. 5(e).

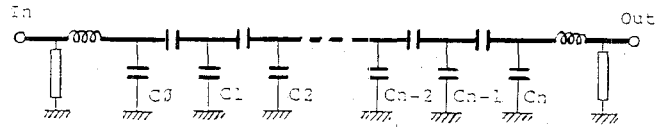
The reason why the configuration of Fig. 2 acts as a filter is easily explained by Fig. 5. Consider the case where the passband of the filter is close to or lower than the resonant frequencies of the resonators. In this case, near the passband Figs. 5(b) and (c) are approximated and simplified to Fig. 5(e); that is, the filter is represented by simple capacitance with respect to the earth. Thus input and output loads can always be matched using simple external matching circuits.

At frequencies lower than the passband, the effect of the series capacitances becomes relatively large, resulting in the stopband of the filter. At frequencies higher than the passband, i.e., at frequencies near the antiresonant frequencies, the filter is represented by Fig. 5(d). Since resonator impedances become large due to the antiresonances, these frequency ranges correspond to the stopband of the filter. At frequencies sufficiently higher than the passband, the filter can be simplified again, as shown in Fig. 5(e). However, since the effect of the capacitance with respect to the earth becomes relatively large, the filter comes into the stopband again.

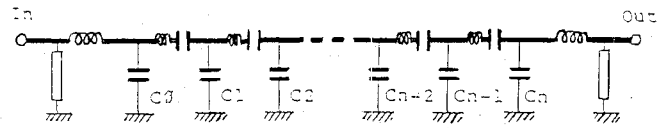
Since the passband is located near the resonant frequencies and the stopband is located near the antiresonant frequencies in the above configuration, this filter is suitable for the duplexer transmitter filter, which must have



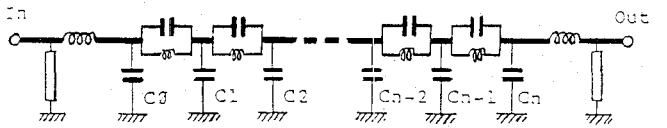
(a)



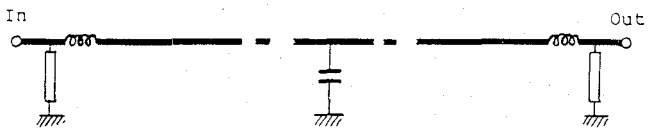
(b)



(c)



(d)



(e)

Fig. 5. Simplified equivalent circuit for Fig. 4's ladder-type filter. (a) For all frequency. (b) Below resonant frequency. (c) Near resonant frequency. (d) Near antiresonant frequency. (e) Above antiresonant frequency.

very sharp cutoff frequency responses at the higher frequency side with respect to the passband.

This configuration also offers both high-power characteristics and very low insertion losses. This is because at the passband of the filter most of the input energy is transferred electrically through the cascade-connected resonators, and the degradation of IDT electrode fingers is very small due to the negligible influence of SAW vibrations. The experiments will illustrate these facts.

IV. ESTIMATION OF CAPACITANCES AGAINST EARTH

One new feature of this developed high-power configuration is the utilization of capacitances against the earth. Capacitors against the earth in Fig. 4's equivalent circuit play a major role in synthesizing steep frequency re-

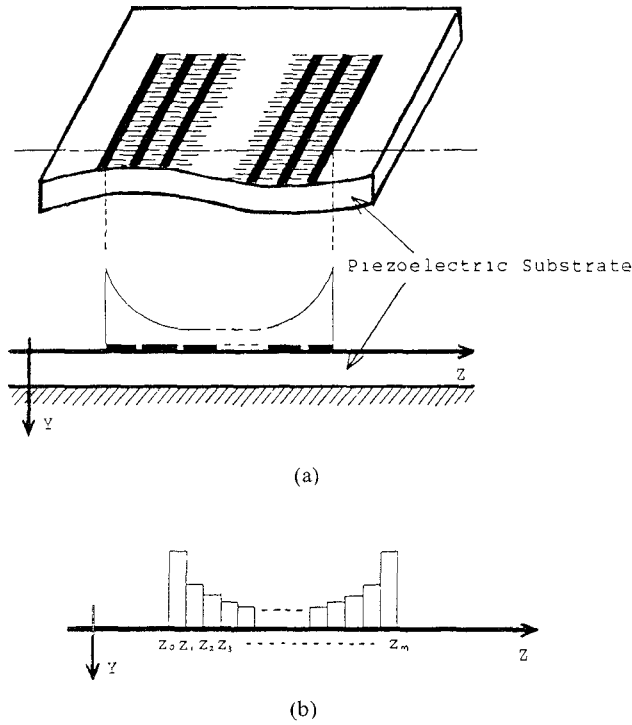


Fig. 6. Filter perpendicular section and charge distribution. (a) Charge distribution on electrode. (b) Step-function approximation.

sponses. Accurate computer simulation results, shown subsequently, will illustrate this point.

It is rather difficult mathematically to formulate these capacitances, because of the complicated filter pattern. However, a fairly precise estimation is possible using some approximation.

The capacitance values for resonator electrode fingers and resonator-resonator coupling pattern areas must be calculated separately. However, the difference for capacitances in both regions is negligible. This is because, if the substrate is sufficiently thick compared with the resonator electrode period, capacitances between electrode fingers and the earth nearly equal the capacitances of the simple parallel plates (see Appendix I). Thus it is not necessary to distinguish the electrode regions from coupling pattern areas.

The filter perpendicular section is illustrated in Fig. 6(a). By identifying the electrode fingers as a uniform metal pattern, each capacitance given in the ladder-type equivalent circuit of Fig. 4 can be estimated using integral equation procedures.

The resonator charge distribution reveals increasing functions toward the edges of the filter due to the fringing effect. This is also shown schematically in Fig. 6(a). A mathematical formulation for capacitances is possible by applying a step-function approximation to each resonator's charge distribution.

From fundamental electrostatic equations [7],

$$\mathbf{E} = -\nabla\phi \quad (1)$$

$$\mathbf{D} = \epsilon \cdot \mathbf{E} \quad (2)$$

$$\nabla \cdot \mathbf{D} = \rho \quad (3)$$

where

\mathbf{E} electric field,

\mathbf{D} electric flux density,

ϕ electric potential,

ρ charge density,

ϵ tensor dielectric constant,

∇ gradient operator.

Using a Fourier series expansion

$$\phi = \sum_{m=0}^{\infty} \tilde{\phi}^m(y) \cos\left(\frac{m\pi}{M}z\right) \quad (4)$$

where \sim represents the Fourier coefficient and M is a large width in the z direction, in which almost all the static field is confined. Putting (4) into (1), (2), and (3) and introducing boundary conditions at $y=0$ and $y=h$, an integral equation is derived which relates charge distribution to potential on the electrodes:

$$V = \int_{z_0}^{z_n} Q(z') G(z|z') dz', \quad z_0 \leq z \leq z_n \quad (5)$$

where V and $Q(z)$ are the potential and the charge distribution on the electrodes. $G(z|z')$ is Green's function:

$$G(z|z') = \frac{h}{\epsilon_{22}M} + \frac{2}{\pi\sqrt{\epsilon_{22}\epsilon_{33}}} \sum_{m=1}^{\infty} \frac{\sinh(\gamma_m h)}{m \cosh(\gamma_m h)} \cdot \cos\left(\frac{m\pi}{M}z\right) \cos\left(\frac{m\pi}{M}z'\right) \quad (6)$$

$$\gamma_m = \frac{m\pi}{M} \sqrt{\frac{\epsilon_{33}}{\epsilon_{22}}}$$

Using the step-function approximation shown in Fig. 6(b), $Q(z)$ can be represented as follows:

$$Q(z) = \sum_{i=1}^n a_i q_i(z) \quad (7)$$

$$q_i(z) = \frac{U(z - z_{i-1}) - U(z - z_i)}{z_i - z_{i-1}}$$

where U is the step function.

Putting (7) into (5) and using Galerkin's method [7] yields algebraic equations. $V=1$ is assumed:

$$1 = \sum_{i=1}^n a_i \int_{z_{i-1}}^{z_i} \int_{z_{j-1}}^{z_j} q_i(z') G(z|z') q_j(z) dz' dz, \quad j=1 \sim n. \quad (8)$$

From (8) capacitors C_i 's in the ladder-type equivalent circuit of Fig. 4 are given as follows:

$$C_0 = \frac{a_1}{2} \quad C_n = \frac{a_n}{2} \quad C_i = \frac{a_i + a_{i+1}}{2} \quad i=1 \sim (n-1). \quad (9)$$

V. COMPUTER SIMULATION PROCEDURES AND EXPERIMENTAL RESULTS

Four parameters are used to design this filter's required frequency responses. These are the resonant frequencies of the resonators, the electrical impedances of the resonators,

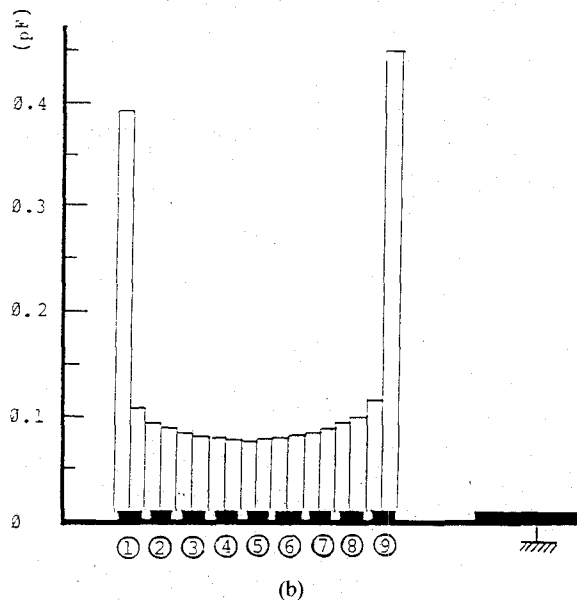
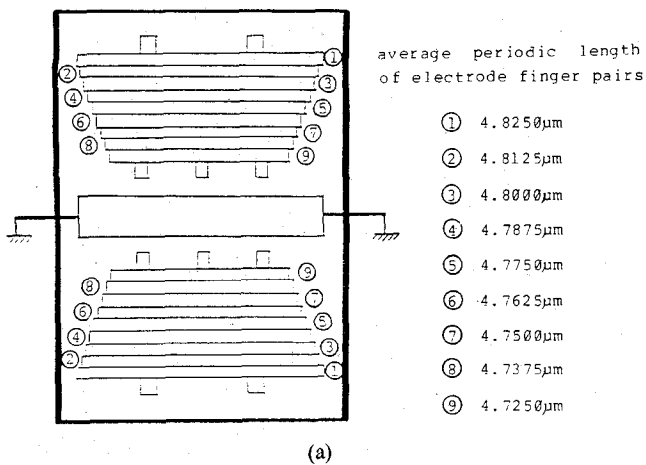


Fig. 7. Schematic drawing and capacitances for high-power SAW filter. (a) Filter pattern. (b) Calculated capacitances.

capacitance values of the capacitors against the earth, and the filter degree corresponding to the number of resonators.

A schematic pattern drawing of the new high-power SAW filter used in computer simulation and experiments is shown in Fig. 7(a). Eighteen series-connected resonators with 400 electrode finger pairs are used. To suppress electric direct coupling between input and output terminals, an isolation earth pattern is introduced in the middle of the filter. The average periodic length of electrode finger pairs for each resonator is also given in Fig. 7(a).

Capacitance values calculated by the procedures shown in Section IV are given in Fig. 7(b). In the calculation, $36^\circ\text{YX-LiTaO}_3$ [4] with 0.35 mm thickness is used as a substrate. Capacitance increases toward the edge of the filter. This increase is more evident toward the center-side edge of the filter due to the influence of the earth pattern.

In general, frequency shifts for high-frequency filters are remarkable due to the influence of floating capacitances. A more exact ladder-type equivalent circuit representation,

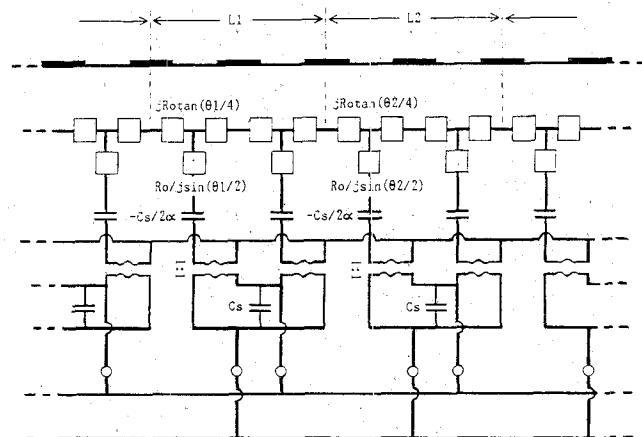


Fig. 8. Equivalent circuit for SAW resonator.

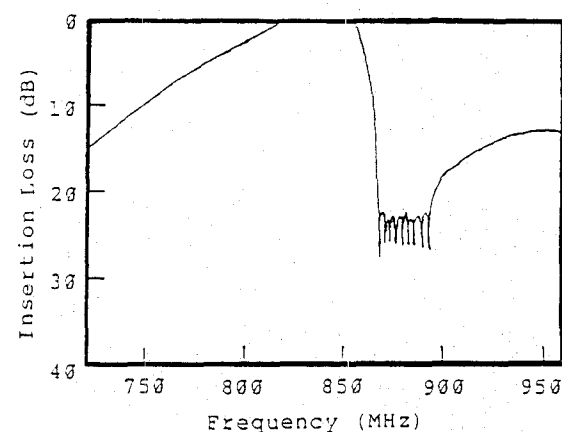


Fig. 9. Simulation results for high-power SAW filter.

which includes floating capacitances between resonators, is discussed in Appendix II.

The previously published equivalent circuit model for the IDT, that is, the mixed field model with α of 0.5 [8], is used to design the SAW resonators. The mixed field model is shown in Fig. 8. Two types of periodic length are introduced in each SAW resonator to achieve the average period given in Fig. 7(a). This is also shown in Fig. 8. A mathematical formulation for the entire SAW resonator is possible using the above equivalent circuit models connected acoustically in cascade and electrically in parallel [9].

Computer simulation results for the new high-power SAW filter using $36^\circ\text{YX-LiTaO}_3$ with 0.35 mm thickness as a substrate are shown in Fig. 9. This simulation uses the more exact ladder-type equivalent circuit including floating capacitances (Appendix II). Simulation results satisfy the required frequency responses for a duplexer transmitter filter and have a rejection level with over 20 dB at receiver band.

Experimental results are shown in Fig. 10. Low insertion loss, under 1.2 dB, is achieved at the transmitter band. The rejection level at the receiver band is over 20 dB. Computer simulation and experimental results agree fairly well.

The chip pattern of the new high-power SAW filter is shown in Fig. 11(a). Chip size is about 2 mm^2 . Electrode

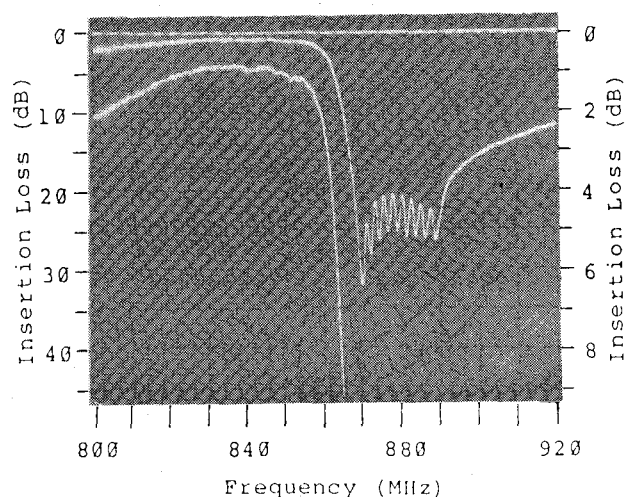
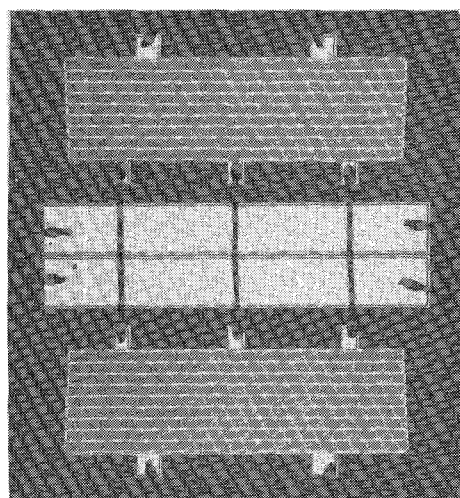
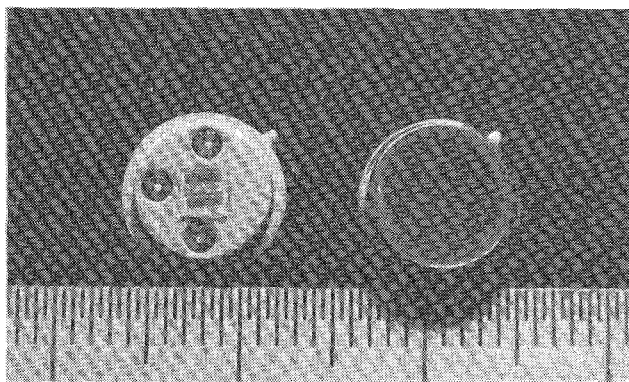


Fig. 10. Experimental results for high-power SAW filter.



(a)



(b)

Fig. 11. Photographs of high-power SAW filter. (a) Chip pattern. (b) Mounted on TO-5 package.

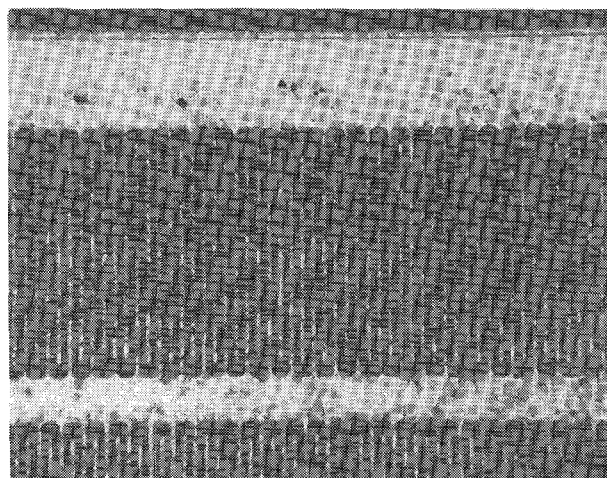
fingers of about $1.2 \mu\text{m}$ width are used in each resonator. The filters mounted on a TO-5 package are also shown in Fig. 11(b). The package on the right is hermetically sealed.

VI. RESULTS OF POWER AGING TESTS

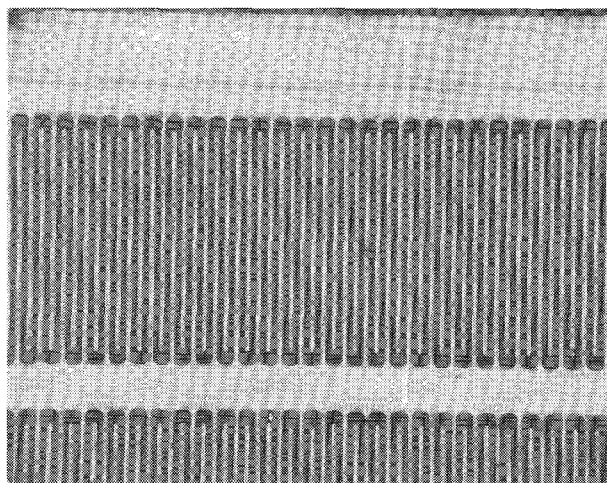
As mentioned above, the maximum input power limit is about 10 ~ 15 dBm at 800 MHz in conventional transversal SAW filters. However, the new filter also provides

TABLE I
AGING TEST FOR HIGH-POWER SAW FILTER

Output Power	Pure Al Electrodes ($1.1\sim 1.2 \mu\text{m}$)	Al-Cu Electrodes ($1.1\sim 1.2 \mu\text{m}$)
2 W	610 hours	no degradation (3000 hours)
3 W	246 hours	no degradation (3000 hours)
4 W	100 hours	no degradation (3000 hours)



(a)



(b)

Fig. 12. Photographs of electrode fingers after aging tests. (a) Pure Al electrodes (2 W output after 600 hours). (b) Al-Cu electrodes (4 W output after 3000 hours).

high-power characteristics. This is because at the passband of the filter, almost all input energy (1.6 W) is electrically transmitted through the connected SAW resonators. The mechanical migration of electrode fingers is very small due to the negligible influence of SAW vibrations. This is very different from previously published low-loss SAW filters [1], [2]. Thus this filter provides high power characteristics.

At the stopband of the filters, strong SAW coupling occurs. However, input energy at this band is very small, such as noise generated from the high power amplifier (HPA). So the degradation of electrode fingers is not remarkable.

System specifications require a heat run test of more than 2000 hours with 2 W output power. Considering the margin, an aging test is conducted with three power levels: 2, 3, and 4 W output power.

It is well known that in both SAW filters and silicon IC's, aluminum electrode fingers doped with 1.5 ~ 2.5 percent copper improve input power by about 5 dB [5], [6]. Thus aluminum-copper electrode fingers have also been used in this high-power SAW filter.

The results of the aging test for the high-power SAW filter are summarized in Table I. The lifetime for pure aluminum electrode fingers with 2 W output power is 610 hours. By contrast, over 3000 hours are obtained even with 4 W output power for aluminum-copper electrode fingers.

Photographs of electrode fingers after aging tests are shown in Fig. 12. The degradation of pure aluminum electrode fingers after a 600 hour, 2 W output power aging test is shown in Fig. 12(a). Mechanical migrations, i.e., many hillocks and voids similar to those of electrical migrations due to high current density, are observed on electrode fingers. Aluminum-copper electrode fingers after a 3000 hour, 4 W output power aging test are shown in Fig. 12(b). No degradation for electrode fingers is observed. These experimental results confirm the feasibility of a SAW antenna duplexer as a class II portable telephone transceiver.

VII. APPLICATION TO ANTENNA DUPLEXER FOR PORTABLE TELEPHONE

As is schematically illustrated in Fig. 1, the newly developed antenna duplexer module includes not only transmitter and receiver filters T1 and R1, but also a second receiver filter R2 and a receiver low-noise amplifier. The authors have previously developed high-performance SAW filters which can be used as receiver filters [1], [2]. They have also developed a bipolar transistor low-noise amplifier using hybrid IC technologies.

Frequency responses for the SAW antenna duplexer module using the developed high-power filter as a transmitter filter and the previously published high-performance SAW filters as receiver filters are shown in Fig. 13. The frequency characteristics required for the antenna duplexer of a portable telephone used in 800 MHz cellular radio systems are satisfied over a wide temperature range, as the figure clearly shows.

Photographs of a miniature SAW antenna duplexer module are shown in Fig. 14. The external appearance of the module is shown on the left. The transmitter high-power SAW filter as well as two receiver high-performance SAW filters on TO-5 packages are fastened by screws. The internal circuits of the module are shown on the right. The circuits also include a low-noise amplifier. The volume of this module is under 8 ml. This is about a one-fifth size reduction over conventional semicoaxial duplexers.

In the near future, the size of transceiver units for cellular radio systems will be greatly reduced, leading to pocket-sized telephones. The fundamental experiments reported herein prove that SAW technologies are suitable for

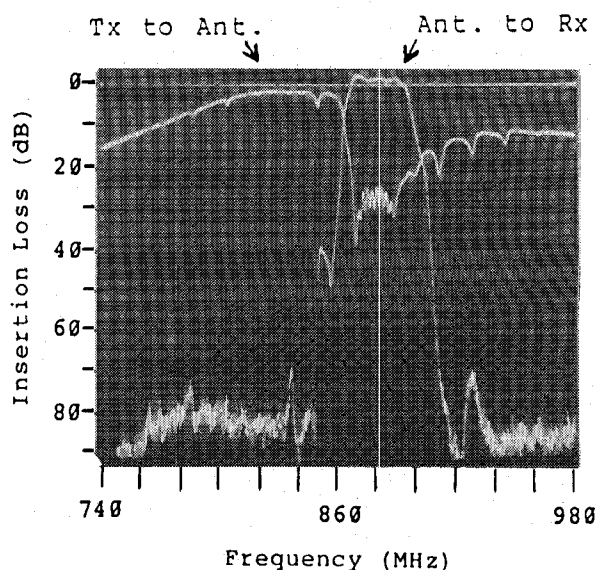


Fig. 13. Frequency characteristics for miniature SAW antenna duplexer module.

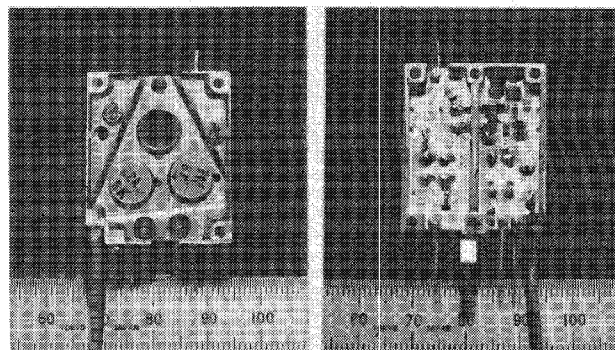


Fig. 14. Photographs of miniature SAW antenna duplexer module.

the miniaturization of transceiver units. New functional devices combining SAW and other active or passive circuit elements, such as low-noise amplifiers and receiver mixers, are now also possible.

VIII. CONCLUSIONS

1) A new ladder-type low-loss, high-power SAW filter has been constructed with wide-band SAW resonators and capacitors between resonator patterns and the earth. It improves the high-power characteristics of conventional transversal SAW filters by over 15 dB.

2) To design the high-power SAW filter, very accurate computer simulation procedures have been developed. Simulation agrees fairly well with experimental results using $36^\circ\text{YX-LiTaO}_3$ as a substrate.

3) Aluminum electrode fingers doped with 1.5 ~ 2.5 percent copper are used to withstand 1.6 W output power. Thus this SAW filter can be used as a class II portable telephone transceiver.

4) A miniature antenna duplexer module including the high-power SAW filter, two high-performance receiver SAW filters, and a low-noise amplifier satisfies the system specifications for an 800-MHz portable telephone.

5) These newly developed technologies show possibilities not only for further size reduction of the duplexer, but also for new functional devices combined with other active and passive circuit elements.

APPENDIX I

Capacitances between electrode fingers and the earth can be estimated by introducing periodic structure approximation, as shown in Fig. 15. As the number of electrode fingers is large enough and the influence of unperiodicity at the edge of the resonator is negligible, these capacitances can be calculated using a conformal mapping technique.

Using a Jacobian elliptic function, the g plane of Fig. 16(a) is transformed into the s plane of Fig. 16(b):

$$s = sn\left(\frac{g}{\alpha}, k_0\right). \quad (A1)$$

Here sn is a Jacobian elliptic function with the following characteristics [10]:

$$sn(K_0, k_0) = 1 \quad (A2)$$

$$sn(K_0 + jK'_0, k_0) = \frac{1}{k_0}. \quad (A3)$$

K_0 is a complete elliptic integral of the first kind with the modulus k_0 . K'_0 is also a complete elliptic integral of the first kind with the modulus $k'_0 = \sqrt{1 - k_0^2}$. From the correspondence of both Figs. 16(a) and (b),

$$\alpha = \frac{L}{2K_0} \quad (A4)$$

$$\frac{K_0}{K'_0} = \frac{rL}{2h}. \quad (A5)$$

The w plane of Fig. 16(c) is also transformed into the t plane of Fig. 16(d):

$$t = A \cdot sn(w, k_1) \quad (A6)$$

$$sn(K_1, k_1) = 1 \quad (A7)$$

$$sn(K_1 + jK'_1, k_1) = \frac{1}{k_1}. \quad (A8)$$

K_1 and K'_1 are complete elliptic integrals of the first kind with moduli k_1 and $k'_1 = \sqrt{1 - k_1^2}$, respectively.

From Fig. 16(b) and (d),

$$A = sn\left(\frac{l}{2\alpha}, k_0\right) \quad (A9)$$

$$\frac{A}{k_1} = \frac{1}{k_0} \quad (A10)$$

$$\frac{k_1}{k_0} = sn\left(\frac{K_0 l}{L}, k_0\right). \quad (A11)$$

Thus using approximate representations for Jacobian elliptic functions [10], capacitance $C = (2\epsilon_{22}K_1)/(rK'_1)$ is

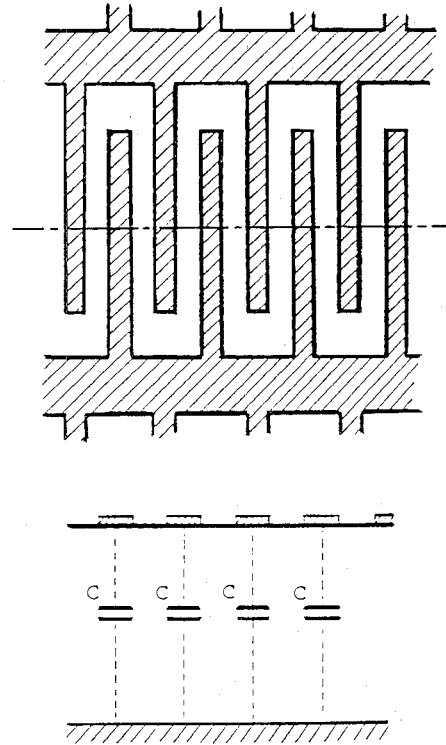


Fig. 15. Capacitances between electrode fingers and the earth.

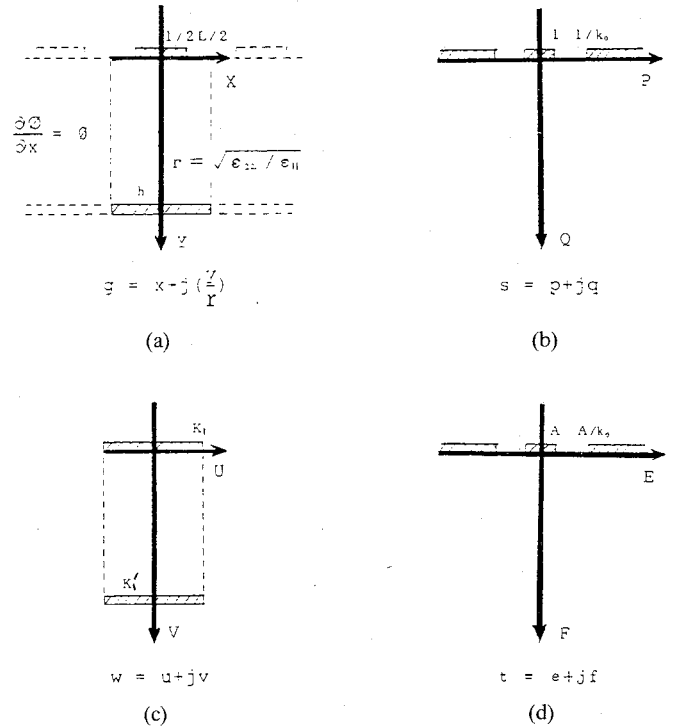


Fig. 16. Conformal mapping for estimation of capacitances. (a) Unit section for electrode. (b) Conformal mapping of (a). (c) Parallel-plate condenser. (d) Conformal mapping of (c).

given as

$$C \approx \frac{\epsilon_{22}L}{h} \cdot \frac{1}{1 - \frac{rL}{\pi h} \ln\left(\sin\left(\frac{\pi l}{2L}\right)\right)}. \quad (A12)$$

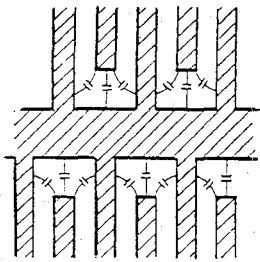


Fig. 17. Electrode finger edge floating capacitances.

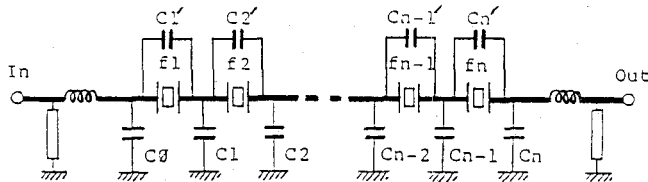


Fig. 18. Ladder-type equivalent circuit including floating capacitances.

Introducing actual parameters ($L = 2.5 \mu\text{m}$, $l = 1.25 \mu\text{m}$, $h = 350 \mu\text{m}$) for the 800-MHz SAW resonators with $36^\circ\text{YX-LiTaO}_3$ substrate into (A12), C is found to be

$$C \approx \frac{\epsilon_{22}L}{h}. \quad (\text{A13})$$

Equation (A13) indicates the capacitance for the uniform parallel-plate condenser. This means it is not necessary to distinguish electrode fingers from coupling pattern areas to estimate capacitances against the earth.

APPENDIX II

The ladder-type equivalent circuit in Fig. 4 is not an exact circuit for the high-power SAW filter in Fig. 2. This is because the influence of floating capacitances is not negligible in high-frequency SAW filters. The electrode finger edge capacitances shown in Fig. 17 are especially remarkable due to the large number of finger pairs. A rather exact ladder-type equivalent circuit representation including these capacitances is given in Fig. 18.

Mathematical formulation for floating capacitors C'_i 's is very difficult because of the complicated electrode finger patterns. Thus C'_i 's are estimated experimentally. In computer simulation of the filter, 3 pF is employed for each C'_i .

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Mitsutaka Hikita (M'87) received the B.S., M.S., and Ph.D. degrees in electronics engineering, all from Hokkaido University, Sapporo, Japan, in 1972, 1974, and 1977, respectively.

In 1978, he joined Central Research Laboratory, Hitachi, Ltd., Tokyo, Japan. From 1972 to 1978, he was engaged in research work on the analysis of electromagnetic-field problems, microwave acoustics, and acoustooptic interactions. From 1978 to 1983, he worked on high-performance SAW filters for radio communication equipment. Recently, he has been engaged in RF system design for cellular radio, as well as advanced application of SAW technologies.

Dr. Hikita is a member of the Institute of Electronics and Communication Engineers of Japan.

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Yoshikatsu Ishida graduated from the electric course of Miyazaki Technical High School, Miyazaki, Japan, in 1963.

He joined Taga Works, Hitachi Ltd., Ibaragi, in 1963. Since joining Tokai Works, Hitachi Ltd., Ibaragi, in 1973, he has been engaged in the development of audio equipment, SAW filters for cellular radio, and SAW filter application devices.



Toyoji Tabuchi graduated from the electric course of Tadotsu Technical High School, Kagawa, Japan, in 1967.

He joined the Central Research Laboratory, Hitachi Ltd., Tokyo, in 1967, and has been engaged in research on laser application devices, liquid crystal devices, and surface acoustic wave devices.

Mr. Tabuchi is a member of the Institute of Electronics and Communication Engineers of Japan.



Kazuhito Kurosawa received the B.S. degree in electronics engineering from Ibaraki University, Ibaraki, Japan, in 1985.

He joined Tokai Works, Hitachi Ltd., in 1985, and has been engaged in development on SAW filters.

Mr. Kurosawa is a member of the Institute of Electronics and Communication Engineers of Japan.
