On Slot-Coupled Microstrip Antennas and Their Applications to CP Operation—Theory and Experiment

M. IRSADI AKSUN, SHUN-LIEN CHUANG, SENIOR MEMBER, IEEE, AND YUEN TZE LO, LIFE FELLOW, IEEE

Abstract—A simple theory based on the cavity model is developed to analyze microstrip antennas excited by a slot in the ground plane. By using an equivalent magnetic current source at the feed, the electric field under the patch is obtained in terms of a set of cavity modes. In particular, the loci of the slot feed location for achieving the circular polarization and the input impedance are computed and found to be in excellent agreement with the experimentally measured results. Simple but surprisingly accurate formulas for slot-fed circularly polarized microstrip antennas are derived and compared with those for probe-fed counterparts.

I. INTRODUCTION

BECAUSE of the increasing interest in the monolithic integration of microstrip antennas with matching networks, amplifiers, and phase shifters, various types of feeding structures have been proposed and discussed [1], [2]. Six basic methods for feeding microstrip antennas and their major advantages and disadvantages are tabulated in Table I [1]. Recently, slot line feed, co-planar waveguide feed and aperture-coupled feed configurations have attracted much attention [3]–[6] because of their suitable geometries for monolithic integration.

The slot-line fed microstrip patch [3] and aperture-coupled microstrip patch [5], [6] have been analyzed using the moment method. Although this approach is more rigorous, it requires a large amount of computation and, most importantly, does not provide formulas for circular polarization (CP) designs. Using the cavity model theory, simple design formulas for probe-fed CP microstrip antennas have been obtained [8], [10], but they can not be directly applied to slot-fed or aperture-fed configurations. This paper is intended to develop a similar theory and to derive similar design formulas for a broad class of electromagnetically excited CP microstrip antennas, including those using co-planar waveguide feed, and aperture-coupled feed.

The cavity model theory [7]–[9], despite its limitation to thin substrates, can provide simple analytical solutions and much physical insight into the design of antennas. The computation time required is generally two or three orders of magnitude smaller than that for a full-wave analysis thus making a computer aided design (CAD) package more useful. This paper also shows that the theory based on this approach can predict the antenna performance which is in excellent agreement with the experimental result. We shall first discuss the design procedure of the slot-fed microstrip antenna for CP operation with design formulas. Next, we present the theory and the derivations of these formulas. Finally, the computed results are presented and compared with the measured results for a few typical examples. In the discussion, a few results for the probe-fed case are also listed for comparison.

II. DESIGN PROCEDURE FOR CP OPERATION

Since CP operation can generally be obtained only for a narrow band of frequencies, it is important to have an accurate theoretical prediction for the dimensions of the patch and the loci of the feeding points. Otherwise, the design of a CP microstrip antenna would require many time-consuming trials either on the experimental bench or on a computer.

For a slot-fed microstrip antenna for which a magnetic current source is assumed, one needs to specify not only the dimensions of the patch and the positions of the slot but also the slot width (w), length (l), and orientation (θ).

For a nearly square patch with dimensions a and b, as shown in Fig. 1 with a slot feed, the resonant wavenumbers \( k_{01} = \pi/a \) and \( k_{10} = \pi/b \) will be very close to each other, so \( Q \) can be assumed to be the same for both modes. Throughout this paper, the wavenumber \( k \) is defined as the effective
TABLE 1
VARIOUS BASIC METHODS FOR EXCITING THE (0, 1) MODE OF A MICROSTRIP ANTENNA

<table>
<thead>
<tr>
<th>Method</th>
<th>Major advantages</th>
<th>Major disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-Probe feed via hole</td>
<td>- no feed line radiation loss</td>
<td>- complicated and costly in fabrication</td>
</tr>
<tr>
<td></td>
<td>- little coupling between patch &amp; line</td>
<td>- difficult to incorporate feed B.C.'s into analysis</td>
</tr>
<tr>
<td></td>
<td>- different value of Z obtainable by choosing feed location</td>
<td></td>
</tr>
<tr>
<td></td>
<td>- Z of thin patch easily predicted</td>
<td></td>
</tr>
<tr>
<td>2-Microstrip-line edge feed</td>
<td>- possible to print antenna and feeding lines in one step</td>
<td>- inflexible in design since both feed and patch are over the same substrate, resulting in possible erratic radiation for mm-waves</td>
</tr>
<tr>
<td>3-Microstrip-line sandwich feed</td>
<td>- flexible in microstrip line and patch design</td>
<td>- two layers of substrates required</td>
</tr>
<tr>
<td></td>
<td></td>
<td>- difficult for integration with active devices and their heat dissipation</td>
</tr>
<tr>
<td>4-Slot line feed</td>
<td>- simple in fabrication</td>
<td>- possible stray radiation from slot</td>
</tr>
<tr>
<td></td>
<td>- easy in integration with active devices</td>
<td>- limited flexibility in large feeding network layout</td>
</tr>
<tr>
<td></td>
<td>- simple for heat dissipation</td>
<td></td>
</tr>
<tr>
<td></td>
<td>- possible to etch patch and slot in one step</td>
<td></td>
</tr>
<tr>
<td>5-Coaxial-waveguide feed</td>
<td>- same as above</td>
<td>- more space required</td>
</tr>
<tr>
<td></td>
<td>- small stray radiation from feed</td>
<td>- less freedom in large feed network design</td>
</tr>
<tr>
<td></td>
<td>- simple transitions to active devices and MMIC's</td>
<td></td>
</tr>
<tr>
<td>6-Aperture coupled feed</td>
<td>- more design freedom (feeding network and patches can be designed separately to a great extent)</td>
<td>- costly and complex</td>
</tr>
<tr>
<td></td>
<td></td>
<td>- more space under ground plane required</td>
</tr>
</tbody>
</table>

wavenumber, which includes all the loss mechanisms, by

\[ k \approx k_0 \sqrt{1 - j f / 2Q} \]  

(1)

provided that \( Q > 1 \) is satisfied. If the excitation frequency is such that \( f_{10} < f < f_{01} \) (assuming \( a > b \)), the fields associated with both of these modes will be strongly excited. The radiated electric field along the \( z \)-axis will then have the following form:

\[ \mathbf{E} = \hat{x}E_x + \hat{y}E_y \]  

(2)

for a probe feed \([10]\), and

\[ E_x \propto \frac{\sin(\pi w \sin \theta / 2a) \sin(\pi l \cos \theta / 2a) \sin(\pi x_0/a)}{k - k_{10}} \]  

(4a)

\[ E_y \propto -\frac{\sin(\pi w \cos \theta / 2b) \sin(\pi l \sin \theta / 2b) \sin(\pi y_0/b)}{k - k_{01}} \]  

(4b)

for an arbitrarily oriented slot feed. Equations (4a) and (4b) will be derived in Section III. Hence using (3a), (3b) and (4a), (4b)

\[ \frac{E_y}{E_x} \equiv \frac{k - k_{10}}{k - k_{01}} = \pm j \]  

(5)

for CP operation, where + (or -) sign is for left (or right)-
hand circularly polarized (LHCP or RHCP) waves, and

\[
A = \begin{cases} 
\cos(\pi y_0/b) \\
\cos(\pi x_0/a)
\end{cases} \quad (6a)
\]
for probe feed

\[
\begin{align*}
\sin(\pi w \cos \theta/2b) \sin(\pi w \sin \theta/2b) \\
\sin(\pi w \sin \theta/2a) \sin(\pi w \cos \theta/2a)
\end{align*}
\]
for slot feed.

(6b)

From the above discussion, we obtain the following design procedure for the CP microstrip patch with both probe feed and slot feed:

1) Choose the frequency of operation \(f\).
2) Choose the effective dimensions \(a\) and \(b\) satisfying

\[
a \geq \frac{c}{2f \sqrt{\varepsilon_r}} \geq b
\]

where \(c = 3 \times 10^8 \text{ m/s}\), and \(\varepsilon_r\) is the relative dielectric constant of the material. The true physical dimensions for \(a\) and \(b\) are smaller by about the thickness of the substrate \(d\) because of the fringing effect.
3) Measure (or estimate) the quality factor, \(Q\).
4) Find \(|A|\) from the following relation [10]:

\[
|A|_1 = \frac{a-b}{b} Q \pm \left[ \left( \frac{a-b}{b} Q \right)^2 - \frac{a}{b} \right]^{1/2}
\]

if the radical is real; otherwise, no CP is possible, and the values of \(a\) and \(b\) should be modified. Here \(|A|_1\) refers to the plus sign and \(|A|_2\) to the minus sign.
5) Determine the feed point \((x_0, y_0)\) by

\[
A = \pm |A|_1
\]

where \(A\) on the left-hand side is given by (6a) or (6b) for a chosen \(\theta\), \(w\) and \(l\), and the positive sign is for LHCP and the negative sign is for RHCP.
6) Occasionally, iterations are needed for fine adjustment.

As an example, the feed loci for a CP microstrip patch with a probe feed and a slot feed are shown in Figs. 2(a) and 2(b), respectively, for two different \(Q\) values. In the case of the slot feed, the locus of the feed is that of the slot center \((x_0, y_0)\).

When \(d \ll \lambda\) the cavity model [7] assumes first that, under the patch, \(E\) is independent of \(z\) and has only a \(z\)-component, namely,

\[
E = z E_z(x, y)
\]

and, second, the patch perimeter can be enclosed by a perfect magnetic wall. Starting from Maxwell’s equations with magnetic current density \(M\), we obtain the following equation:

\[
\nabla^2 E_z + k^2 E_z = \mathbf{\hat{z}} \cdot (\nabla \times M)
\]

and the equation for its associated Green’s function

\[
\nabla^2 G_z + k^2 G_z = \delta(x - x') \delta(y - y'),
\]

Equation (12) can be solved in terms of a set of modal func-
tions, \( \{ \Psi_{mn}(x, y) \} \), satisfying
\[
\nabla^2 \Psi_{mn} + k_{mn}^2 \Psi_{mn} = 0
\]
(13a)
\[
\frac{\partial \Psi_{mn}}{\partial n} = 0, \quad \text{on the magnetic wall}
\]
(13b)
where \( n \) is a unit outward normal vector to the magnetic wall.

The Green’s function can then be expressed as
\[
G_z(x, y; x', y') = \sum_{m,n=0}^{\infty} \frac{\Psi_{mn}(x', y') \Psi_{mn}(x, y)}{k^2 - k_{mn}^2}.
\]
(14)
Hence, the field distribution due to an arbitrarily oriented slot can be found by the convolution integral
\[
E_z(x, y) = \int dxdy' G_z(x, y; x', y') 
\]
\[
\cdot [\hat{z} \cdot \nabla' \times M(x', y')].
\]
(15)
Since the substrate thickness \( d \) is typically a few hundredths of a wavelength, the magnetic current in the slot may also be assumed to be independent of \( z \) [11]. The validity of this approximation is justified by the agreement between experiment and theory. Then the magnetic current density, for simplicity in computation, is approximated in an average sense as
\[
M = \hat{u} \left[ U \left( \frac{u + \frac{w_e}{2}}{2} \right) - U \left( \frac{u - \frac{w_e}{2}}{2} \right) \right]
\]
\[
\cdot \left[ U \left( u + \frac{l_e}{2} \right) - U \left( u - \frac{l_e}{2} \right) \right] / w_e l_e
\]
(16)
in the transformed coordinate system, \( u \) and \( v \), as shown in Fig. 1, where \( w_e \) and \( l_e \) are the effective width and length of the slot, respectively, and \( U \) is a unit step function. It should be noted that this approximation of the magnetic current density is good for small slots. Using (16), \( E_z \) is obtained from (15):
\[
E_z(x, y) = \sum_{m,n=0}^{\infty} \frac{f_{mn}(x_0, y_0)}{k^2 - k_{mn}^2} A_{mn}
\]
\[
\cdot \cos \left( \frac{m\pi}{a} x \right) \cos \left( \frac{n\pi}{b} y \right)
\]
(17a)

where
\[
f_{mn}(x_0, y_0) = \frac{A_{mn}}{w_e} \left\{ \sin \left( \frac{m\pi}{a} x_0 + \frac{n\pi}{b} y_0 \right) \right. \\
\left. \cdot \sin \left( \frac{m\pi}{a} \sin \theta + \frac{n\pi}{b} \cos \theta \right) \right\}
\]
\[
\times \sin \left( \frac{m\pi}{a} \cos \theta + \frac{n\pi}{b} \sin \theta \right) \left( \frac{l_e}{2} \right)
\]
\[
\cdot \sin \left( \frac{m\pi}{a} \frac{x_0}{a} - \frac{n\pi}{b} \frac{y_0}{b} \right) \\
\times \sin \left( \frac{m\pi}{a} \sin \theta + \frac{n\pi}{b} \cos \theta \right) \left( \frac{w_e}{2} \right)
\]
\[
\cdot \sin \left( \frac{m\pi}{a} \cos \theta - \frac{n\pi}{b} \sin \theta \right) \left( \frac{l_e}{2} \right).
\]
(17b)

\section*{A. Far-Field Components}

When the observation point \( r \) is very far from the source region \( r' \), the far-field components are approximated as
\[
E(r) \approx \frac{j \omega \mu e^{-j k r}}{4 \pi r} \hat{r} \times \int \int dS' e^{j k r' r} M_e(r')
\]
(18a)
\[
H(r) \approx (I - \hat{r} \hat{r}') \cdot \frac{-j \omega \mu e^{-j k r}}{4 \pi r} \int \int dS' e^{j k r' r} M_e(r')
\]
(18b)
for a surface magnetic current density \( M_e \) [12], where \( I \) is the unit dyad. In our case, the equivalent surface magnetic current density can be approximately found from
\[
M_e = -2 \hat{u} \times \hat{z} E_z
\]
(19)
where \( E_z(x, y) \) given by (17a) is evaluated for a typical point on the magnetic wall. Finally, the far-field components are obtained by substituting (19) into (18a) and (18b), where, for simplicity, the free-space Green’s function is used. Thus the substrate effect outside the patch is ignored. A multilayered dyadic Green’s function [13] for an infinite structure could be used if one needs to compute approximately the surface wave loss. In general, it does not necessarily give a better result since all practical antennas are of finite dimensions. Anyhow, it is believed that the error due to the use of free-space Green’s function on the calculations of impedance and radiation pattern should be negligible for thin substrate.

\section*{B. Input Impedance}

The input impedance of an arbitrarily oriented slot feed can be found by the power-voltage relation as
\[
\frac{1}{Z_{in}} = \frac{P^*}{|V|^2} = -\int M^* \cdot H d\tau
\]
(20a)
\[
= \frac{1}{j \omega \mu d} \left[ \int_{-\infty}^{\infty} \sum_{m,n=0}^{\infty} \frac{f_{mn}(x_0, y_0)}{k^2 - k_{mn}^2} \frac{l_e}{w_e} \right]
\]
(20b)
where \( V \approx M d w_e \) is used. Equation (20a) is the dual of the electric probe feed case [7]. Since this equation has a double infinite summation, the computation converges relatively slow. But for two special cases, \( \theta = 0^\circ \) and \( \theta = 90^\circ \), the input impedance can be cast into a form which has only one infinite summation over \( m \) and \( n \), respectively, and the computation takes much less time.

The input admittance (20b) can be interpreted as a network consisting of an inductance in parallel with an infinite number of series resonant circuits, each corresponding to the modes of the cavity. The inductive term, \( l_e/j \omega \mu \), is due to the divergenceless part of the \( H \) field. In other words, it is related to the conservation law of magnetic charge [14]. Hence the admittance can be obtained by substituting the magnetic field \( H \)
\[
H = -\nabla \times E - \frac{M}{j \omega \mu},
\]
(21)
into the definition of the input admittance, (20a),
\[
\frac{1}{Z_m} = \frac{\int \mathbf{M} \cdot \nabla \times \mathbf{E} \, d\tau}{j\omega\mu|V|^2} + \frac{\int \mathbf{M} \cdot \mathbf{M} \, d\tau}{j\omega\mu|V|^2}.
\] (22)

The second term yields the inductive term, \( I_e/j\omega\mu \, d\omega_e \), after substituting \( \mathbf{M} \) from (16) and \( V \approx M\omega_e \).

The assumption of a unity uniform magnetic current density through the thickness of the substrate may cause stronger excitation than the actual case, unless the intensity is properly corrected. The effective length and width of the slot are used to compensate for this problem as demonstrated in the next section. Also note that the slot should be fed at its center for the input impedance measurement.

C. CP Operation

For the TM\(_{10}\) mode, the surface magnetic current density is given by
\[
\mathbf{M}_0 = -2\mathbf{a} \times \mathbf{E}_z = -2A_{10} f_{10}(x_0, y_0) \frac{k^2 - k_{10}^2}{k^2 - k_{10}^2}, \quad \text{at } x = 0, a.
\] (23)

When substituting (23) into (18a) the far field in the z-direction is obtained
\[
\mathbf{E}(r) \approx \frac{jk_0 e^{-jkr}}{4\pi r} 4A_{10} bd f_{10}(x_0, y_0) \frac{k^2 - k_{10}^2}{k^2 - k_{10}^2}. \] (24)

Similarly, for the TM\(_{01}\) mode
\[
\mathbf{M}_0 = 2A_{01} f_{01}(x_0, y_0) \frac{k^2 - k_{01}^2}{k^2 - k_{01}^2}, \quad \text{at } y = 0, b.
\] (25)

\[
\mathbf{E}(r) \approx \frac{jk_0 e^{-jkr}}{4\pi r} 4A_{01} ad f_{01}(x_0, y_0) \frac{k^2 - k_{01}^2}{k^2 - k_{01}^2}. \] (26)

Using the approximation \( k^2 - k_{01}^2 \approx 2k_{01}(k - k_{01}) \), one obtains (4), (5), and (6b). Following [8] and [10], there exist, in general, two solutions of \( A \), as given in (8), in terms of the Q-factor of the patch. Let the two solutions be \(|A_1| \) and \(|A_2|\); then their corresponding CP frequencies are
\[
f_1 = f_{10} \frac{1}{1 + \frac{|A_1|}{2Q}} = f_{01} \frac{1 + |A_1|}{2Q} \] (27a)
\[
f_2 = f_{10} \frac{1}{1 + \frac{|A_2|}{2Q}} = f_{01} \frac{1 + |A_2|}{2Q} \] (27b)

Usually, \(|A_1|, |A_2| \approx 1 + c/b \approx 1\). Since the positive sign is used for the definition of \(|A| \) in (8), then \(|A_1| \geq 1 \) whereas \(|A_2| \leq 1 \), which makes \( f_1 \) less than or equal to \( f_2 \). The case \( f_1 = f_2 \) is possible only when \(|A_1| = |A_2| \), i.e., (8) has a double root solution. Moreover, the order of frequencies \( f_{10} < f_1 < f_2 < f_{01} \) always holds, as can be seen from (27a) and (27b).

From the previous analysis, we see that the excitation of the CP microstrip antennas depends on (a) the frequency of excitation \( f_1 \) or \( f_2 \), (b) the slot location and (c) the slot orientation angle \( \theta \). The frequency of excitation is a function of the quality factor \( Q \), in such a way that \( f_1 \) and \( f_2 \) get closer to \( f_{10} \) and \( f_{01} \), respectively, with the increase of \( Q \). The last two factors, (b) and (c), control the degree of coupling between the magnetic current and the cavity modal fields as can be seen from the product \( \mathbf{M}^* \cdot \mathbf{H} \) in (20a). As an example, there will be no excitation of the TM\(_{01}\) mode by a vertically oriented magnetic current density, \( i.e., \theta = 90^\circ \), and no excitation of the TM\(_{10}\) mode by a horizontally oriented magnetic current density, \( i.e., \theta = 0^\circ \). In general, the angle \( \theta \) determines the relative strength of the two modal fields. As \( |\theta| \) decreases, the TM\(_{01}\) modal field increases relative to that of TM\(_{10}\) mode, and vice versa.

From (5), (6b), and (8), it is not difficult to explain the qualitative behavior of the loci of the slot locations for CP excitation, and from the relation \(|A_1|/|A_2| \approx 1 \) and (6b), to explain the symmetries of the loci as shown in Figs. 2 and 3. It should be noted that the solution for feed location close to the edges should not be taken seriously, due to the edge effect.
Fig. 4. Input impedance versus frequency plot of a slot-fed rectangular microstrip antenna. $\varepsilon_r = 2.62$, $d = 3.4$ cm, $b = 3.0$ cm, $w = 0.07$ cm, $l = 0.7$ cm, $d = 0.0794$ cm, and $\theta = 0^\circ$.

Fig. 5. Input impedance versus frequency plot of a slot-fed rectangular microstrip antenna. $\varepsilon_r = 2.62$, $a = 3.4$ cm, $b = 3.0$ cm, $w = 0.1$ cm, $l = 1.0$ cm, $d = 0.3175$ cm, and $\theta = 0^\circ$.

IV. RESULTS AND DISCUSSIONS

Measurements and theoretical calculations have been carried out for several slot-fed rectangular microstrip antennas, all made of copper-clad Rexolite 2200 with a relative permittivity of 2.62. Here, the emphasis is on the input impedance and the CP operation because of their importance in applications. The other features of these antennas, such as radiation patterns, multislots, feeds, and dual frequency operations, will be discussed elsewhere [15].

Figs. 4 and 5 are the plots of the input impedance of the slot-fed patch antenna with 1/32 and 1/8 in thick substrates, respectively. For comparison purposes, both computed and measured results are shown. For both cases, the slots were oriented horizontally, $\theta = 0^\circ$; thus, only the TM$_{01}$ mode was excited. The physical dimensions of the slots and patches are $a = 3.4$ cm, $b = 3.0$ cm for $d = 0.0794$ cm, $w = 0.07$ cm, and $l = 0.7$ cm for Fig. 6; and 2) $d = 0.3175$ cm, $w = 0.1$ cm, and $l = 1.0$ cm for Fig. 5. As can be seen from Fig. 5, this theory predicted the input impedances very accurately even for the substrate as thick as 0.05$\lambda$. In computation, actually the effective widths and lengths for 1) $d = 0.0794$ cm, $w_e = 0.04$ cm, $l_e = 0.69$ cm, and 2) $d = 0.3175$ cm, $w_e = 0.03$ cm, $l_e = 0.5$ cm are used. As the thickness of the substrate increases, the effective slot dimensions decrease. This is expected since we used a uniform magnetic current density model in (16) throughout the substrate thickness but

the actual structure has the slot in the ground plane. The excitation, according to this model, would be too strong if the same dimensions were used. Fig. 6 shows the computed and measured input impedances for the slot angle $\theta = 47^\circ$. In all of the examples, the shapes of the locus and the frequency distribution on the impedance locus are in good agreement with those for the experiments. The close agreement in these examples supports the usefulness of the theory. The convergence in the computation of input impedances for the horizontal (or vertical) slot-fed patches is very fast because of the availability of a closed-form expression for one of the double infinite series in the input impedance expression (20). Unfortunately, for an arbitrarily oriented slot, no closed-form expression was found and computation time increased. Nevertheless, as noted before, this theory is still much more efficient computationally than the full-wave analysis. For all the cases considered here, we used the length extension by about a value of $d$ for $a$ and $b$ to account for the fringing field effect.

Based on the theory discussed in Section III and the design procedures given in Section II, several experimental patch antennas were designed for CP operation with various substrate thicknesses. A typical result for a RHCP microstrip antenna on a 1/8 in thick substrate is shown in Fig. 7. For this experiment, an excitation cable was placed across the slot at its center, as shown in the inset of Fig. 7. The measured CP radiation pattern is shown in that same figure. But this is not the only way to excite the patch through the slot, an aperture-coupled feed, co-planar waveguide feed, or any other feed structure using a slot can also be used. An aperture-coupled feed and a co-planar waveguide-fed CP microstrip antennas were also designed following the same procedure and using the same dimensions as given above [16]. We have obtained similar CP operation.

V. CONCLUSION

A simple and efficient method has been presented to analyze electromagnetically coupled microstrip antennas with good accuracy. The method is based on the cavity model, and can be applied not only to the rectangular patches but to patches of many other geometries as well. The feed loci for CP operation and the input impedance of several slot-fed antennas are calculated and tested with experiments. The close agreement
Fig. 7. Radiation pattern for RHCP microstrip antenna on a 0.3175 cm thick substrate. $a = 3.15$ cm, $b = 3.0$ cm, $w = 0.1$ cm, $l = 0.5$ cm, $\theta = 30^\circ$, $x_0 = 0.9$ cm, and $y_0 = 1.0$ cm.

between the experiments and theoretical calculations proves the usefulness of the theory.

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REFERENCES


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Yuen Tze Lo (S’49–A’53–M’58–SM’66–F’69–LF’86), for a photograph and biography please see page 958 of the August 1989 issue of this TRANSACTIONS.