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II. Reciprocity in Antenna Theory

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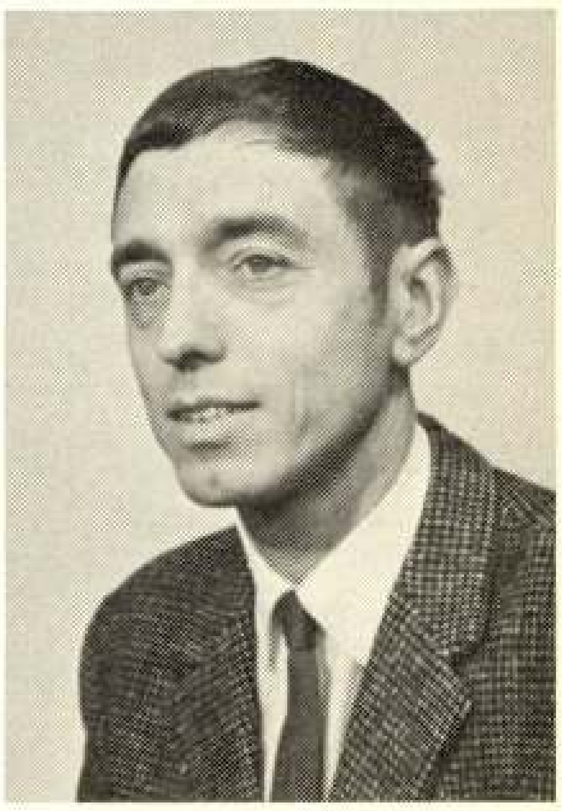

Synopsis: A reciprocity relation between the transmitting and the receiving properties of an antenna is given. In the transmitting situation a certain part of the antenna, called 'source domain', is capable of carrying external currents, both of the electric and the magnetic type. In the receiving situation a plane electromagnetic wave is incident upon the antenna system.

In particular, the reciprocity relation is applied to a microwave antenna system in which the feeding waveguide operates in more than a single mode.

1. Introduction

One of the basic theorems in the electromagnetic theory of antennas is a reciprocity relation between the transmitting and the receiving properties of an antenna. The customary form of this reciprocity relation (see, for example, [1]) applies to two different antennas, a finite distance apart, each of them playing in turn the role of transmitting antenna or receiving antenna. In technical applications, however, one frequently employs the reciprocity theorem to relate the radiation pattern of a transmitting antenna to the pattern of the same antenna when it is receiving an incident plane wave. Clearly, for the latter problem a reciprocity theorem applicable to a single antenna is required. Now, such a reciprocity theorem can be obtained from the aforementioned reciprocity theorem applicable to two separate antennas by letting the mutual distance between the antennas become very large and considering the limiting form of the relation between the electromagnetic field quantities involved. However, a reciprocity theorem applicable to a single antenna can also be obtained directly, without having to resort to the corresponding problem for two separate antennas; this has been shown by the author in an earlier paper [2].

In the present paper the relevant reciprocity theorem, and the conditions under which it holds, are briefly stated, whereupon the application to a microwave antenna system is worked out in detail.

2. Description of the configuration

The antenna system under consideration occupies a bounded domain V in space; the boundary of V is a sufficiently regular closed surface S. The Cartesian co-ordinates of a point in space are denoted by x, y and z; the time variable is denoted by t. The position vector is denoted by $\mathbf{r}$. The electromagnetic fields occurring in the transmitting as well as in the receiving situation are assumed to vary sinusoidally in time with the same angular

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frequency ω. The complex representation of the field vectors is used; in the formulas the complex time factor $\exp(-i\omega t)$, common to all field components, is omitted.

The antenna consists of a (partly lossy) medium whose electromagnetic behaviour is assumed to be linear. The properties of the medium may suddenly change when crossing a (bounded) surface; across such a discontinuity surface the tangential parts of both the electric and the magnetic field vector are continuous. Other parts of the antenna system may consist of conducting surfaces. These surfaces are assumed to be electrically perfectly conducting; on them the tangential part of the electric field vector vanishes.

In the transmitting situation a subdomain V_{source} of V is capable of carrying 'external' currents, both of electric and magnetic type. These external currents represent the sources through which power of nonelectromagnetic origin can be delivered to the system. The boundary of V_{source} is a sufficiently regular closed surface S_{source}. In the receiving situation no external currents are present in the antenna system.

The medium outside the antenna system is assumed to be linear, homogeneous, isotropic, and lossless (which includes the case of free space) with real scalar permittivity ε_0 and real scalar permeability μ_0.

In the following **E**, **H**, **D** and **B** denote the space and frequency dependent complex representations of the electric field vector, the magnetic field vector, the electric flux density and the magnetic flux density, respectively. The superscripts 'T' and 'R' are used to denote the transmitting and the receiving situation, respectively.

3. The antenna as a transmitting system

In the transmitting situation the antenna carries time harmonic external currents. Let $\mathbf{J}^{\mathrm{T}}$ be the volume density of external electric currents and let $\mathbf{K}^{\mathrm{T}}$ be the volume density of external magnetic currents; $\mathbf{J}^{\mathrm{T}}$ and $\mathbf{K}^{\mathrm{T}}$ differ from zero in V_{source} only. In V the electromagnetic field vectors satisfy the inhomogeneous Maxwell equations:

$$\operatorname{rot} \mathbf{H}^{\mathrm{T}} + i\omega \mathbf{D}^{\mathrm{T}} = \mathbf{J}^{\mathrm{T}}, \tag{3.1}$$

$$\operatorname{rot} \mathbf{E}^{\mathrm{T}} - i\omega \mathbf{B}^{\mathrm{T}} = -\mathbf{K}^{\mathrm{T}} \tag{3.2}$$

and the constitutive equations which express $(\mathbf{D}^{\mathrm{T}}, \mathbf{B}^{\mathrm{T}})$ linearly in terms of $(\mathbf{E}^{\mathrm{T}}, \mathbf{H}^{\mathrm{T}})$. Further, the tangential parts of both $\mathbf{E}^{\mathrm{T}}$ and $\mathbf{H}^{\mathrm{T}}$ are continuous across a surface of discontinuity in material properties. In the domain outside the antenna system $\mathbf{E}^{\mathrm{T}}$ and $\mathbf{H}^{\mathrm{T}}$ satisfy the homogeneous Maxwell equations:

$$\operatorname{rot} \mathbf{H}^{\mathrm{T}} + i\omega\varepsilon_0 \mathbf{E}^{\mathrm{T}} = \mathbf{0}, \tag{3.3}$$

$$\operatorname{rot} \mathbf{E}^{\mathrm{T}} - i\omega\mu_0 \mathbf{H}^{\mathrm{T}} = \mathbf{0}. \tag{3.4}$$

In addition, the transmitted field satisfies the radiation conditions (Cf. [3]). As a consequence the following representation holds at large distances from the antenna system:

$$\mathbf{E}^{\mathrm{T}}(\mathbf{r}_{\mathrm{P}}) \simeq \mathbf{F}^{\mathrm{T}}(\hat{\mathbf{r}}_{\mathrm{P}}) \exp(ik_0 r_{\mathrm{P}})/4\pi r_{\mathrm{P}}, \tag{3.5}$$

$$\mathbf{H}^{\mathrm{T}}(\mathbf{r}_{\mathrm{P}}) \simeq (\varepsilon_0/\mu_0)^{\frac{1}{2}} (\hat{\mathbf{r}}_{\mathrm{P}} \times \mathbf{F}^{\mathrm{T}}) \exp(ik_0 r_{\mathrm{P}})/4\pi r_{\mathrm{P}}, \tag{3.6}$$

in which:

$$k_0 \stackrel{\text{def}}{=} \omega\, (\varepsilon_0\mu_0)^{\frac{1}{2}} = 2\pi/\lambda_0, \tag{3.7}$$

where λ_0 is the wavelength, $\hat{\mathbf{r}}_{\mathrm{P}} \stackrel{\text{def}}{=} \mathbf{r}_{\mathrm{P}}/r_{\mathrm{P}}$ is the unit vector in the direction of observation and r_{P} is the distance from the origin of the coordinate system to the point of observation P. The vectorial amplitude radiation characteristic $\mathbf{F}^{\mathrm{T}}$ of the antenna system only depends upon the direction of observation $\hat{\mathbf{r}}_{\mathrm{P}}$ and is transverse with respect to the direction of propagation of the expanding spherical wave generated by the antenna, i.e. $\hat{\mathbf{r}}_{\mathrm{p}} \cdot \mathbf{F}^{\mathrm{T}} = 0$. Further, $\mathbf{F}^{\mathrm{T}}$ can be expressed in terms of the value the tangential parts of $\mathbf{E}^{\mathrm{T}}$ and $\mathbf{H}^{\mathrm{T}}$ have on the boundary surface S of the antenna system (for details, see [2]).

4. The antenna as a receiving system

In the receiving situation a time harmonic plane electromagnetic wave is incident upon the antenna system. Since in the receiving situation the antenna carries no external currents, the electromagnetic field vectors satisfy in V the homogeneous Maxwell equations:

$$\operatorname{rot} \mathbf{H}^{\mathrm{R}} + i\omega \mathbf{D}^{\mathrm{R}} = \mathbf{0}, \tag{4.1}$$

$$\operatorname{rot} \mathbf{E}^{\mathrm{R}} - i\omega \mathbf{B}^{\mathrm{R}} = \mathbf{0}, \tag{4.2}$$

and the constitutive equations which express $(\mathbf{D}^{\mathrm{R}}, \mathbf{B}^{\mathrm{R}})$ linearly in terms of $(\mathbf{E}^{\mathrm{R}}, \mathbf{H}^{\mathrm{R}})$. Further, the tangential parts of both $\mathbf{E}^{\mathrm{R}}$ and $\mathbf{H}^{\mathrm{R}}$ are continuous across a surface of discontinuity in material properties. In the domain outside the antenna system the scattered field $(\mathbf{E}^{\mathrm{s}}, \mathbf{H}^{\mathrm{s}})$ is introduced as the difference between the (actual) total field $(\mathbf{E}^{\mathrm{R}}, \mathbf{H}^{\mathrm{R}})$ and the incident field $(\mathbf{E}^{\mathrm{i}}, \mathbf{H}^{\mathrm{i}})$, the latter field being the electromagnetic field that would be present if the antenna system were absent:

$$\mathbf{E}^{\mathrm{s}} \stackrel{\text{def}}{=} \mathbf{E}^{\mathrm{R}} - \mathbf{E}^{\mathrm{i}}, \quad \mathbf{H}^{\mathrm{s}} \stackrel{\text{def}}{=} \mathbf{H}^{\mathrm{R}} - \mathbf{H}^{\mathrm{i}}. \tag{4.3}$$

The incident field is the plane wave:

$$\mathbf{E}^{\mathrm{i}} = \mathbf{B} \exp(-ik_0\hat{\boldsymbol{\beta}} \cdot \mathbf{r}), \tag{4.4}$$

$$\mathbf{H}^{\mathrm{i}} = (\varepsilon_0/\mu_0)^{\frac{1}{2}} (\mathbf{B} \times \hat{\boldsymbol{\beta}}) \exp(-ik_0\hat{\boldsymbol{\beta}} \cdot \mathbf{r}), \tag{4.5}$$

where **B** specifies the amplitude and the state of polarization (in general, elliptic) and $-\hat{\boldsymbol{\beta}}$ is the unit vector in the direction of propagation. Since the wave is transverse, we have $\mathbf{B} \cdot \hat{\boldsymbol{\beta}} = 0$. In the domain outside the antenna system both the incident and the scattered field satisfy the homogeneous Maxwell equations:

$$\operatorname{rot} \mathbf{H}^{\mathrm{i,s}} + i\omega\varepsilon_0 \mathbf{E}^{\mathrm{i,s}} = \mathbf{0}, \tag{4.6}$$

$$\operatorname{rot} \mathbf{E}^{\mathrm{i,s}} - i\omega\mu_0 \mathbf{H}^{\mathrm{i,s}} = \mathbf{0}. \tag{4.7}$$

In addition, the scattered field satisfies the radiation conditions (Cf. [3]).

5. The reciprocity relation

The reciprocity relation is obtained by an application of Lorentz's reciprocity theorem of electromagnetic fields [2]. This theorem can be applied, provided that the electromagnetic properties of the medium present in the transmitting situation and those of the medium present in the receiving situation are interrelated in such a way that at all points in space the relation:

$$\mathbf{E}^{\mathrm{R}} \cdot \mathbf{D}^{\mathrm{T}} - \mathbf{E}^{\mathrm{T}} \cdot \mathbf{D}^{\mathrm{R}} - \mathbf{H}^{\mathrm{R}} \cdot \mathbf{B}^{\mathrm{T}} + \mathbf{H}^{\mathrm{T}} \cdot \mathbf{B}^{\mathrm{R}} = 0 \tag{5.1}$$

holds. In the domain outside the antenna system this is obviously the case, as the constitutive equations are here simply $\mathbf{D} = \varepsilon_0\mathbf{E}$ and $\mathbf{B} = \mu_0\mathbf{H}$, both in the transmitting and in the receiving situation. In the antenna system the situation may be more complicated. Equation (5.1) holds *without change* of the properties of the medium if the medium is *reciprocal*. In all other cases the medium is non-reciprocal, and a change in properties has to be made when switching from transmission to reception, and vice versa. When, for example, unidirectional devices are present, the direction of blocking is to be reversed. It is noted that in the general theorem non-reciprocal media as well as media showing the magneto-electric effect [4] are included.

Employing the integral representation of $\mathbf{F}^{\mathrm{T}}$ indicated in Section 3, application of the reciprocity theorem to the domain outside the antenna system leads to:

$$\iint_S(\mathbf{E}^{\mathrm{T}} \times \mathbf{H}^{\mathrm{R}} - \mathbf{E}^{\mathrm{R}} \times \mathbf{H}^{\mathrm{T}}) \cdot \mathbf{n}\mathrm{d}A = (\mathrm{i}\omega\mu_0)^{-1}\mathbf{B} \cdot \mathbf{F}^{\mathrm{T}}(\hat{\beta}), \quad (5.2)$$

where $\mathbf{n}$ denotes the unit vector in the direction of the outward normal to S. Application of the reciprocity theorem to the domain V occupied by the antenna system leads to:

$$\iint_S(\mathbf{E}^{\mathrm{T}} \times \mathbf{H}^{\mathrm{R}} - \mathbf{E}^{\mathrm{R}} \times \mathbf{H}^{\mathrm{T}}) \cdot \mathbf{n}\mathrm{d}A =$$

$$= \iiint_{V_{\mathrm{source}}} (\mathbf{J}^{\mathrm{T}} \cdot \mathbf{E}^{\mathrm{R}} - \mathbf{K}^{\mathrm{T}} \cdot \mathbf{H}^{\mathrm{R}})\mathrm{d}V. \quad (5.3)$$

As the left-hand sides of (5.2) and (5.3) are equal, these equations can be combined to:

$$\iiint_{V_{\mathrm{source}}} (\mathbf{J}^{\mathrm{T}} \cdot \mathbf{E}^{\mathrm{R}} - \mathbf{K}^{\mathrm{T}} \cdot \mathbf{H}^{\mathrm{R}})\mathrm{d}V = (\mathrm{i}\omega\mu_0)^{-1}\mathbf{B} \cdot \mathbf{F}^{\mathrm{T}}(\hat{\beta}). \quad (5.4)$$

Details of the proofs are given in [2].

The reciprocity relation for a particular antenna system at hand can be obtained from either (5.2), (5.3) or (5.4). Which one of these equations is to be selected as a starting point, depends mainly on where the information concerning the antenna properties is needed. In the next section this will be illustrated for a microwave antenna system where the field quantities are measured in a uniform section of the cylindrical waveguide feeding the antenna system (see also [5]).

6. Application to a microwave antenna system

In this section we discuss the application of the reciprocity theorem of Section 5 to the microwave antenna system depicted in Fig. 1. In the transmitting situation the domain indicated as the generator is capable of carrying external currents, in the receiving situation this part of the system acts as the load. The waveguide connecting the generator/load domain to the other parts of the antenna system (amongst which are the reflector or horn and the radiating aperture) is assumed to have a uniformly cylindrical, lossless section. Let M be the (finite) number of propagating modes in this section; all modes of higher order than M are evanescent.

To allow for multimode operation*) of the waveguide, we take $M \geqq 1$. It is further assumed that in the uniform section a transverse reference plane can be chosen, such that with sufficient accuracy the total electromagnetic field in this cross-section can be written as the superposition of the contributions from the propagating modes only. (This implies that the reference plane is sufficiently far removed from non-uniformities in the waveguide.) Let the Cartesain co-ordinate system be chosen such that the plane $z = L$ coincides with the reference plane and the positive z-direction be chosen away from the generator/load. Then the transverse parts of the electromagnetic field vectors in the reference plane can be written as:

$$\mathbf{E}_{\mathrm{transverse}}(x,y,L) =$$

$$= \Sigma_{m=1}^{M}[A_m \exp(\mathrm{i}\beta_m L) + B_m \exp(-\mathrm{i}\beta_m L)]\mathbf{e}_m(x,y), \quad (6.1)$$

$$\mathbf{H}_{\mathrm{transverse}}(x,y,L) =$$

$$= \Sigma_{m=1}^{M}[A_m \exp(\mathrm{i}\beta_m L) - B_m \exp(-\mathrm{i}\beta_m L)]\mathbf{h}_m(x,y), \quad (6.2)$$

where A_m is the complex amplitude of the m-th mode propagating in the positive z-direction, B_m is the complex amplitude of the m-th mode in the negative z-direction and β_m, $\mathbf{e}_m$ and $\mathbf{h}_m$ are the phase factor, the transverse electric field distribution and the transverse magnetic field distribution, respectively, of the m-th mode propagating in the positive z-direction. The transverse mode distributions satisfy the orthogonality relation:

$$\iint_D(\mathbf{e}_m \times \mathbf{h}_n) \cdot \mathbf{i}_z\mathrm{d}x\mathrm{d}y = 0 \text{ if } m \neq n, \quad (6.3)$$

where D is the cross-section of the waveguide and $\mathbf{i}_z$ is the unit vector in the positive z-direction. Further, the normalization constant N_m of the m-th mode is introduced as:

$$N_m \stackrel{\mathrm{def}}{=} \iint_D(\mathbf{e}_m \times \mathbf{h}_m) \cdot \mathbf{i}_z\mathrm{d}x\mathrm{d}y. \quad (6.4)$$

In the *transmitting situation* we write:

$$B_m^{\mathrm{T}} = \rho_m^{\mathrm{T}} A_m^{\mathrm{T}}, \quad (6.5)$$

*) The question as to multimode operation has been raised by Dr. M. E. J. Jeuken, Department of Electrical Engineering, Eindhoven University of Technology, Eindhoven, The Netherlands.

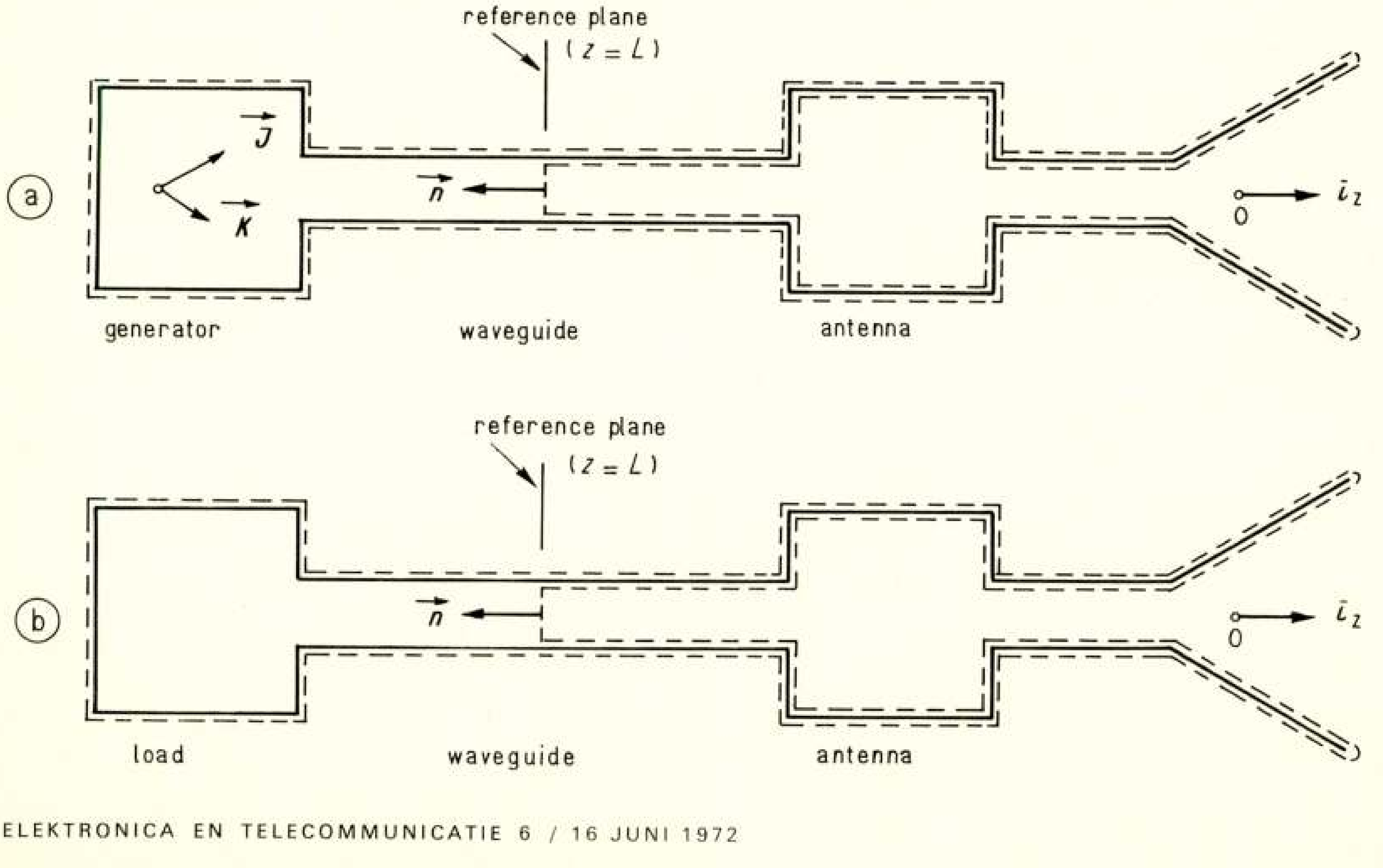

Fig. 1. Reciprocity relation for a microwave antenna system. In the transmitting situation (a) the generator domain is capable of carrying external currents, in the receiving situation (b) a plane wave is incident upon the system.

where ρ_m^T is the reflection factor of the m-th mode in the transmitting situation. This factor is a measure for the mismatch of the radiating part of the antenna with respect to the m-th waveguide mode; if the relevant part of the antenna system is matched to the m-th waveguide mode we have $\rho_m^T = 0$. Due to the linearity of the system we can further write:

$$\mathbf{F}^T(\hat{\mathbf{r}}_P) = \Sigma_{m=1}^M A_m^T \mathbf{F}_m^T(\hat{\mathbf{r}}_P), \tag{6.6}$$

where $\mathbf{F}_m^T(\hat{\mathbf{r}}_P)$ is the contribution from the m-th mode of unit amplitude to the vectorial amplitude radiation characteristic of the antenna in the transmitting situation.

In the *receiving situation* we write:

$$A_m^R = \rho_m^R B_m^R, \tag{6.7}$$

where ρ_m^R is the reflection factor of the m-th mode in the receiving situation. This factor is a measure for the mismatch of the load with respect to the m-th waveguide mode; if the load is matched to the m-th waveguide mode we have $\rho_m^R = 0$.

We now apply (5.2) to the closed surface S indicated by the dotted lines in Fig. 1. Since the waveguide walls, all other walls of the antenna system and the reflector or horn are assumed to be electrically perfectly conducting, we have $\mathbf{n} \times \mathbf{E}^{T,R} = \mathbf{0}$ on these parts. Hence:

$$\int\int_S (\mathbf{E}^T \times \mathbf{H}^R - \mathbf{E}^R \times \mathbf{H}^T) \cdot \mathbf{n} dA =$$
$$= -\int\int_D (\mathbf{E}^T \times \mathbf{H}^R - \mathbf{E}^R \times \mathbf{H}^T) \cdot \mathbf{i}_z dx dy. \tag{6.8}$$

Substituting the expansions (6.1) and (6.2) in the right-hand side, taking into account the orthogonality relation (6.3) and the normalization condition (6.4) and using (6.5), (6.6) and (6.7), we obtain:

$$2\Sigma_{m=1}^M (1 - \rho_m^T \rho_m^R) A_m^T B_m^R N_m = (i\omega\mu_0)^{-1} \mathbf{B} \cdot \Sigma_{m=1}^M A_m^T \mathbf{F}_m^T(\hat{\boldsymbol{\beta}}). \tag{6.9}$$

As (6.9) must hold for arbitrary values of $A_1^T, \ldots, A_M^T$ we have:

$$2(1 - \rho_m^T \rho_m^R) B_m^R N_m = (i\omega\mu_0)^{-1} \mathbf{B} \cdot \mathbf{F}_m^T(\hat{\boldsymbol{\beta}}) \tag{6.10}$$
$$(m = 1, \ldots, M).$$

By measuring ρ_m^T, ρ_m^R and B_m^R and calculating N_m we can determine $\mathbf{B} \cdot \mathbf{F}_m^T(\hat{\boldsymbol{\beta}})$ from (6.10). By varying the direction of propagation $\hat{\boldsymbol{\beta}}$ of the incident plane wave and its state of polarization we then can determine all components of $\mathbf{F}_m^T(\hat{\boldsymbol{\beta}})$ for all relevant directions (Cf. [5]).

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III. Active Receiving Antennas

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Synopsis: An active antenna is a combination of a passive antenna with electronic circuitry. This circuitry can be an amplifier or a frequency converter. The antenna with integrated amplifier is sometimes called 'antennafier', the antenna with integrated converter is an 'antennaverter' [1]. While a passive antenna is reciprocal and can be used as transmitting antenna and as receiving antenna without any change in design, the active circuitry is nonreciprocal and, therefore, active transmitting antennas and active receiving antennas have different design and different operational aspects.

Active antennas offer a wide variety of applications and in many cases improved performance compared to conventional passive antennas. The limited length of this paper demands to discuss only a few characteristic problems which are found with active antennas. Most existing practical experience concerns receiving antennas and integrated transistoramplifiers [5, 6, 7, 8]. Therefore, the receiving antennafier will be the main object of this article.

1. More bandwidth or smaller antennas

A blockscheme of a receiving system is given in Fig. 1: A passive antenna A receives the signal out of the surrounding space and so forms the signal source which feeds the receiving system. This signal source is connected to a passive input network N_1 which is followed by an active electronic element E_1 which is again followed by a passive network N_2 and an electronic element E_2 and so on. The passive networks are used for matching impedances and for filtering unwanted frequencies. The input of the receiving system (consisting of A, N_1 and E_1) plays an important role as far as bandwidth, selectivity, linearity and signal-to-noise ratio is concerned. Optimizing this input is the fundamental problem of active receiving antennas.

For a simple theory of active antennas this input circuitry

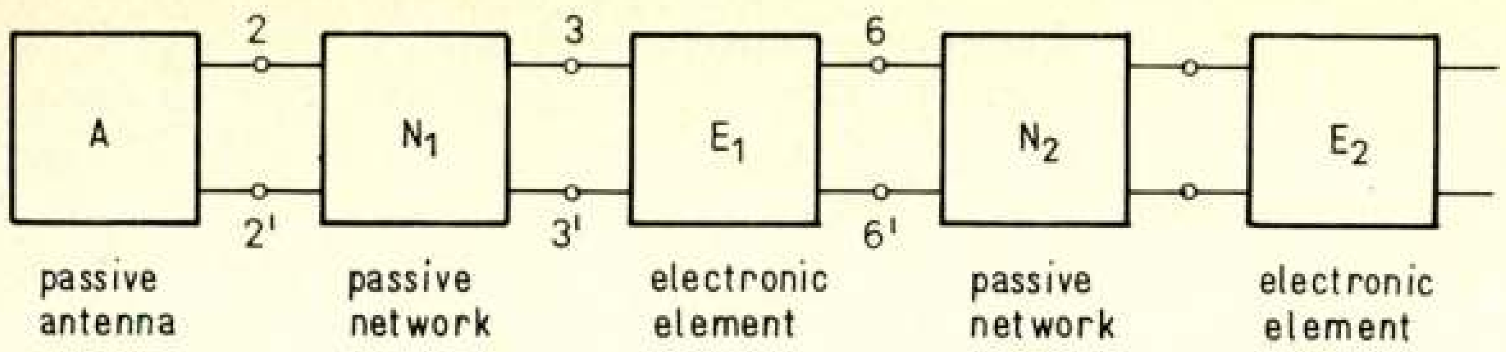

Fig. 1. Scheme of a receiving system.

should be presented as shown in Fig. 2. The passive antenna A is a signal source with ohmic radiation resistance R_1 and an additional twoport between terminals 1-1′ and 2-2′, which represents the inductive and capacitive elements and the losses of the antenna structure. This twoport can be influenced by antenna design. The transistor E_1 can be seen in a simplified manner as the combination of an ideal amplifier and input resistance R_4, a passive input network and a passive output network. The input network of the transistor between terminals 3-3′ and 4-4′ incorporates the input capacity of the transistor, the inductance of the leads and lead resistances. Input bandwidth, input selectivity and signal-to-noise ratio depend on the total passive input network N* between terminals 1-1′ and 4-4′ which includes the passive antenna-structure, the network N_1 of Fig. 1 and the passive transistor input.

input also contains filter-circuitry, which may be tuned or untuned. In this article only the untuned case is considered with the aim of finding the maximum possible bandwidth of the input system. The general problem is not, how to get smaller bandwidth, this is solved by adding filters. In the tuned case this maximum possible bandwidth can be seen as determining the possible tuning range.

The maximum possible bandwidth of the input network N* (between terminals 1-1′ and 4-4′) is determined by the reactive power in its reactive components and by its ohmic terminations R_1 and R_4. The maximum bandwidth is obtained if the network N* has a minimum number of reactive elements and minimum reactive currents. This bandwidth does not only depend on the radiation resistance R_1 and the reactive components of the passive antenna but also on the reactive power in

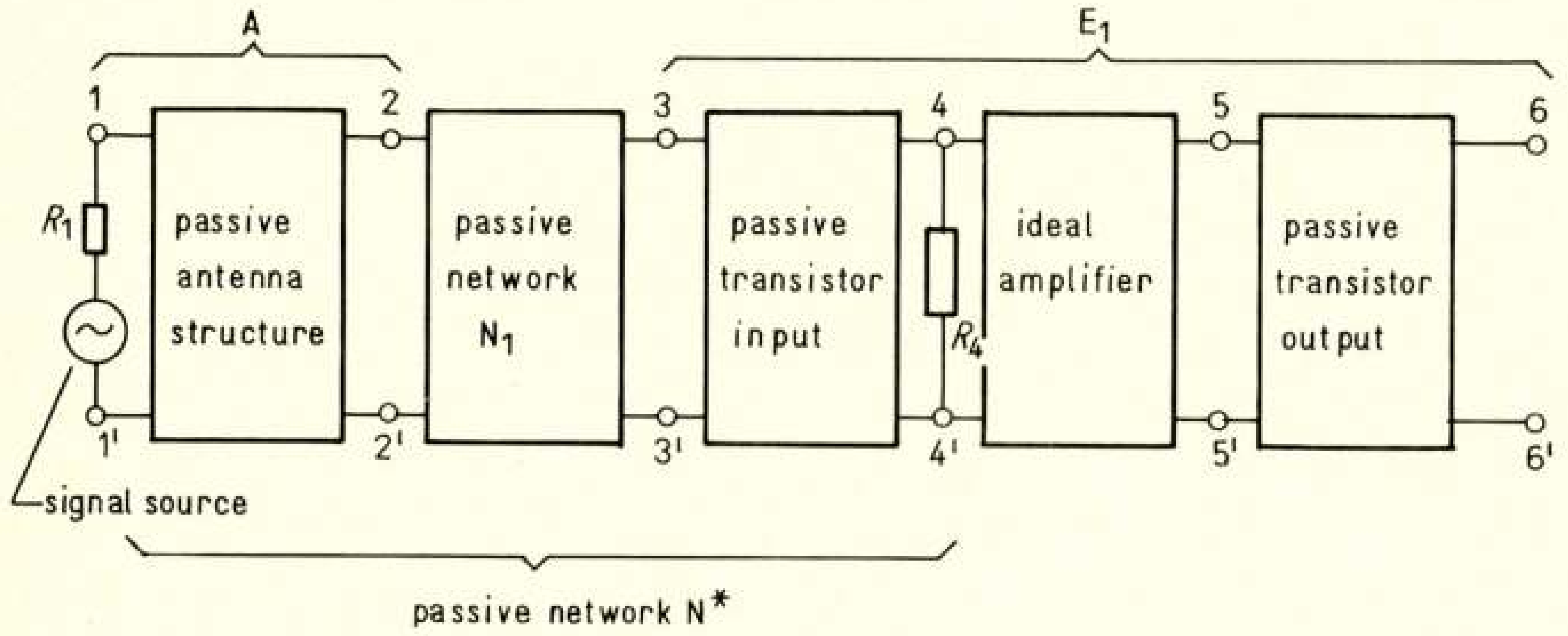

Fig. 2. Passive input network N*.

Fig. 3 shows different forms of this passive network N*. In Fig. 3a the conventional case is shown in which transistor E_1 is part of a receiver, and the passive antenna is connected to the receiver by a line. The network N_1 starts with a matching network to match the antenna impedance (found between terminals 2-2′) to the line. At the end of this line, inside the receiver, there is a second matching network to match the line to the input of the electronic element E_1. Usually, the receiver

the matching networks which again depend on the characteristic impedance of the line section.

To get more input bandwidth the reactive elements are reduced by shifting the line section from N_1 to N_2 (Fig. 1). Then the network of Fig. 3b is obtained which contains only one matching network. This simplified matching can be a fundamental improvement of the input network and gives far more bandwidth. In an optimized technology all parts of Fig. 3b are

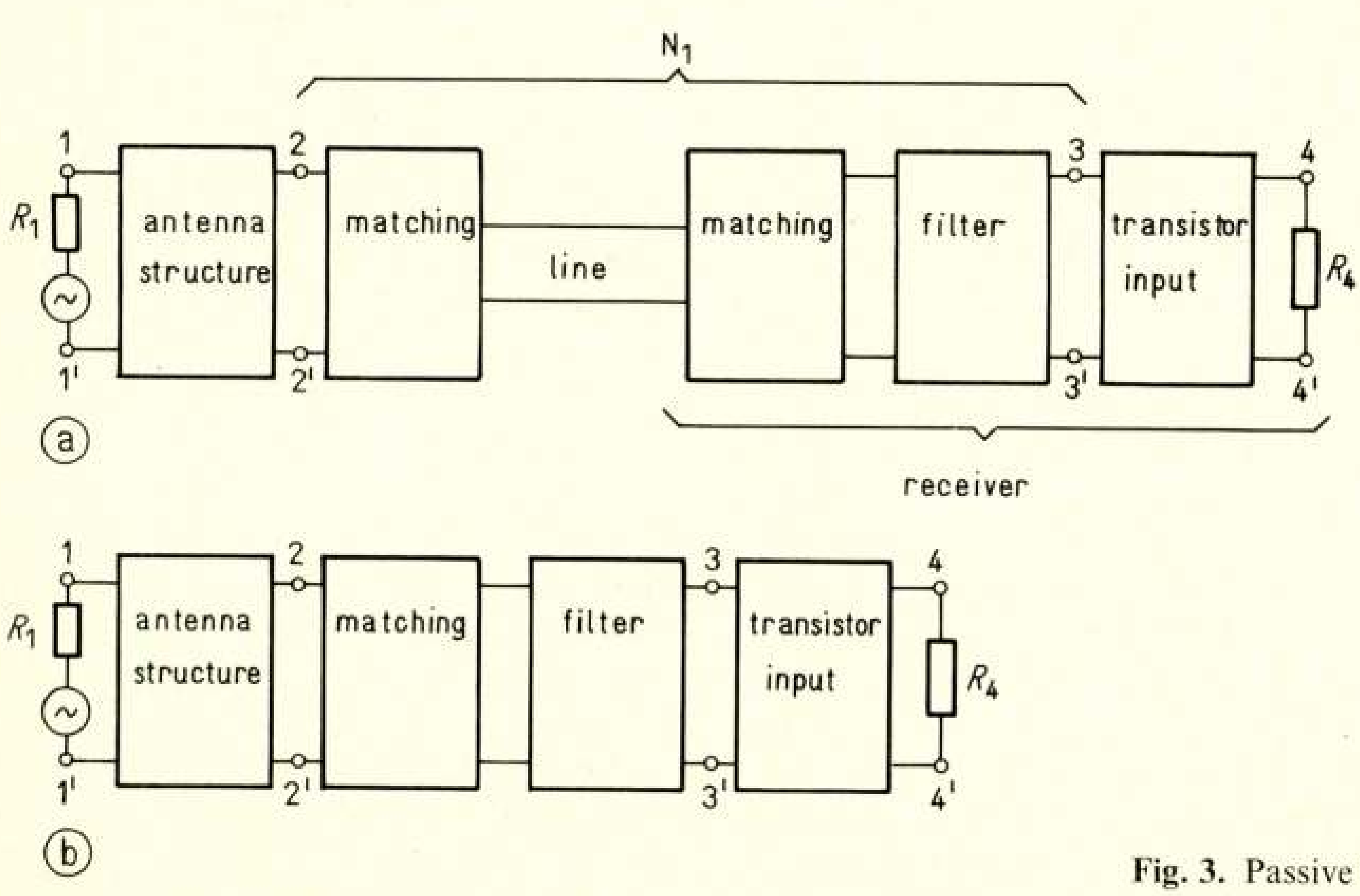

Fig. 3. Passive input network; (a) with and (b) without time section.

integrated with the antenna structure in which the passive transistor-input elements are also parts of the matching and filtering circuitry. So one gets a minimum of circuitry which gives maximum possible bandwidth for a given transistor type and for a given antenna size. Such an integrated combination is called an active antenna. Fig. 4 shows an example of an integrated dipole

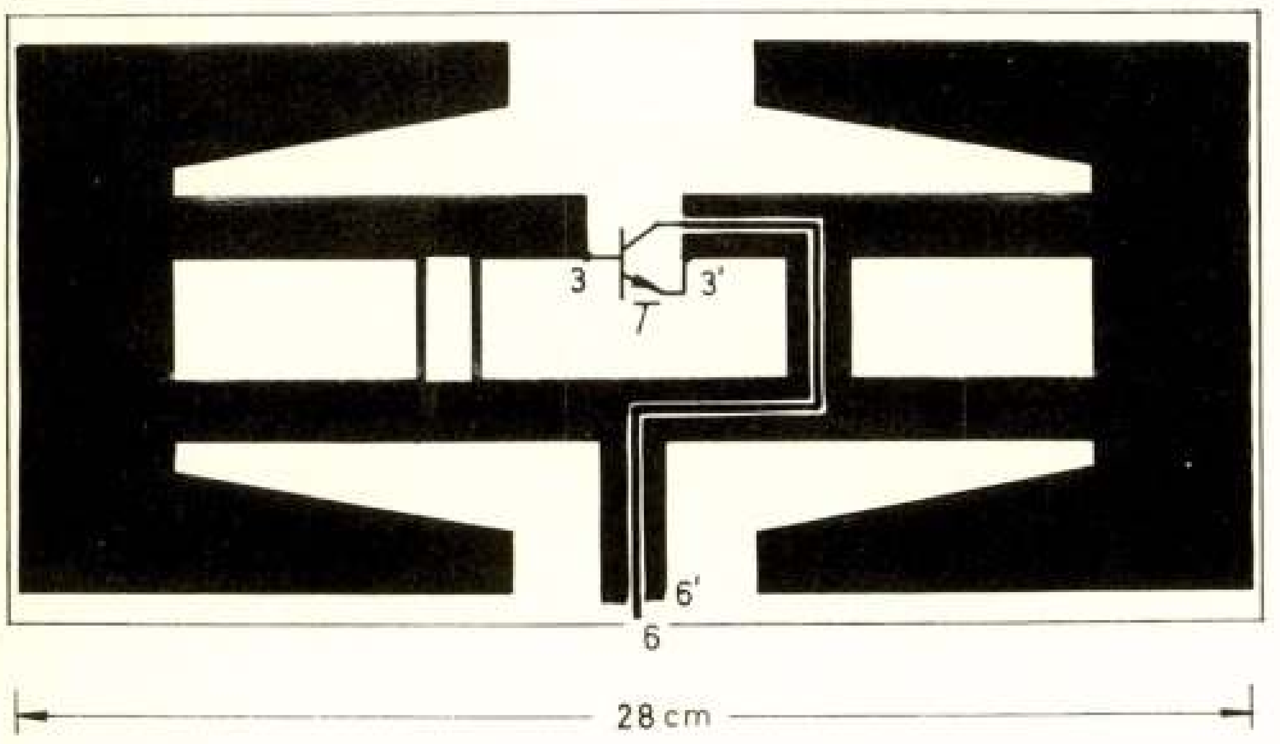

Fig. 4. Active integrated dipole.

antenna in printed-circuit technology for a frequency band from 190 ... 230 MHz [2]. The design of Fig. 4 includes the transistor-output with a transition from a symmetrical dipole to a coaxial line at terminals 6-6'. For lower frequencies and for miniature antenna structures printed coils are used as inductances, as is shown in Fig. 5 for a monopole [3, 6].

Fig. 5. Printed antenna inductance.

Of course, the ohmic components R_1 and R_4 play an important part in forming bandwidth. As is well known, the radiation resistance R_1 of dipoles or of monopoles decreases with decreasing antenna length. Consequently, the maximum possible input-bandwidth decreases with decreasing antenna length. In most of the practical cases there is a prescribed input-bandwidth. Therefore, for prescribed bandwidth the integrated minimized antenna must have a special length which generally is smaller than the length of conventional antennas for equal input-bandwidth. This reduction in antenna size can be very significant in many cases.

Besides, the active antennas have a very simple technology together with well defined impedances of antenna and network and in the circuitry of Fig. 3b a more accurate matching than in the more complicated circuitry of Fig. 3a. Furthermore, the broadband matching of the line section in Fig. 3a is never ideal and causes some reflection at both ends so that matching depends on line length and is somewhat undefined.

In the case of receiving antennas the power level is so low that the input circuitry, if wanted, can be tuned by means of varactor diodes [3, 4]. The number of tuning diodes necessary for a given tuning range decreases with decreasing reactive power in the input network. So the integrated minimized network can be tuned with a minimum number of tuning diodes.

2. Improved signal-to-noise ratio

A second important feature of active receiving antennas is the improved signal-to-noise ratio of the receiving system. For a receiving antenna the output power is not an important quantity because nearly unlimited amplification can be introduced in the receiver. So the active antenna must primarily be designed to obtain the optimum signal-to-noise ratio for the input system.

The noise in the receiving system of Fig. 1 has three main sources:

- the signal-source noise which is received by the passive antenna out of the surrounding space;
- the noise of the passive network N* due to its losses;
- the noise of the amplifying element E_1.

If E_1 gives sufficient amplification the noise of N_2 and E_2 and further elements in Fig. 1 can be neglected [9].

The influence of the network losses on the signal-to-noise ratio has two different aspects: These losses lower the signal amplitude at the transistor input 4-4' and so lower the ratio of signal-amplitude to transistor noise. Furthermore, the losses are caused by the ohmic components of N*, which are additional thermal noise sources.

To optimize the system one has to minimize losses in the network N*. This can be done by minimizing the reactive power in the network, by using a low-loss antenna structure and a transistor with low-loss input circuitry. So it is evident that the integrated antenna by its lower losses also gives an improved signal-to-noise ratio compared to conventional systems.

This integrated antenna offers an additional improvement in signal-to-noise ratio by using noise match. This will be explained here for the case of a lossless input network N*. The total noise of a receiving system can be expressed by a system noise-temperature T_s [9, 10], also called operational noise-temperature [11]. If there is lossless network N* the system noise-temperature T_s is the sum of antenna noise-temperature T_A and the equivalent noise-temperature T_T of transistor E_1:

$$T_s = T_A + T_T. \tag{1}$$

T_A describes the noise which the antenna receives out of the surrounding space together with the signal. This antenna noise is a collection of components due to man-made noise, atmospheric discharges, thermal noise of the environment and cosmic radiation. The antenna noise-temperature T_A depends on frequency and is time-variable. Standard values of T_A are known from many international measurements [12] and are shown in Fig. 6. T_A has very high values at low frequencies and decreases with increasing frequency asymptotically to 300 K. It may be pointed out that the values of T_A in Fig. 6 are standard values

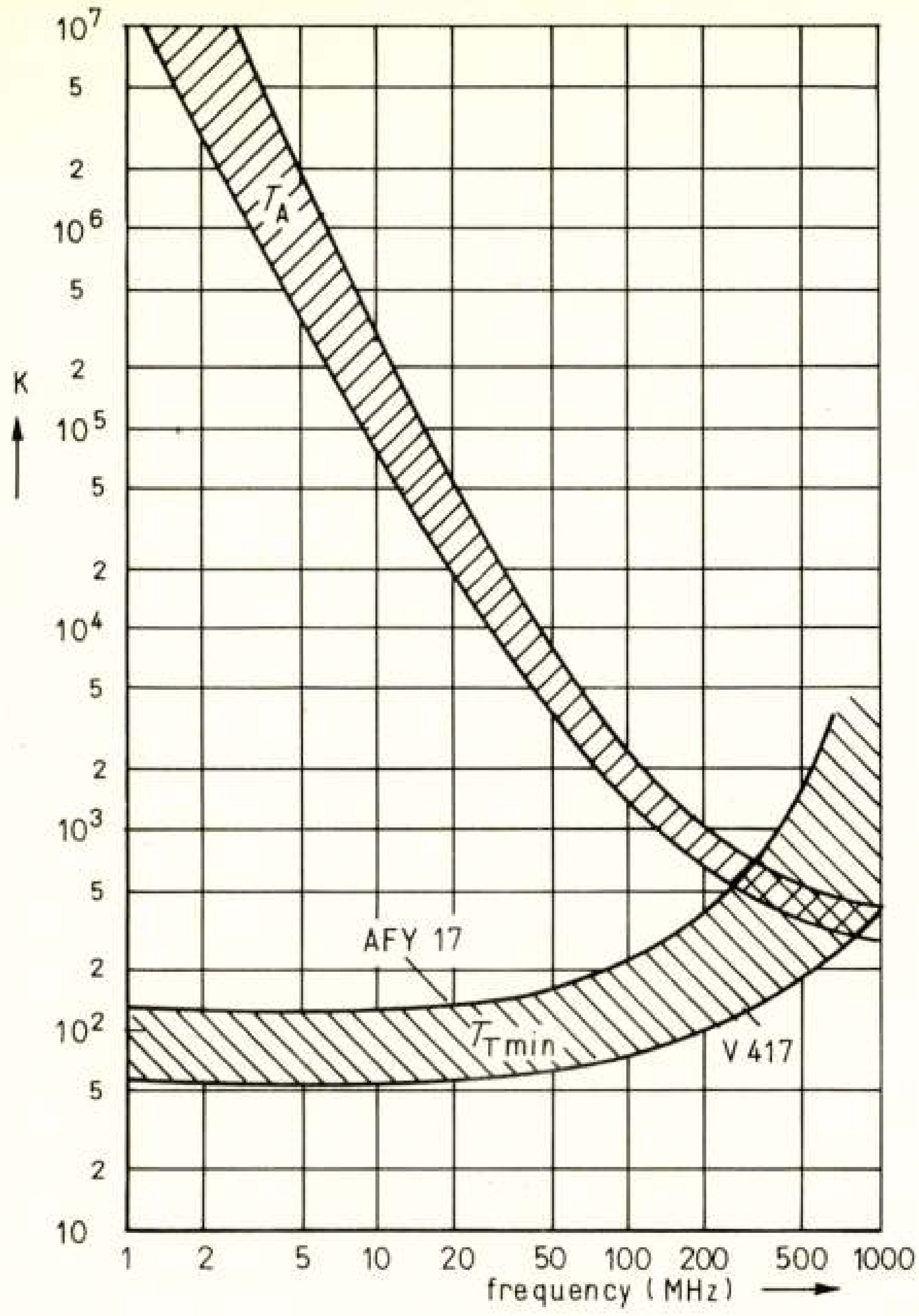

Fig. 6. Antenna noise temperature T_A and minimum transistor-noise-temperature T_{Tmin}.

which can be used successfully for antenna development. Locally T_A may deviate considerably from this in some cases.

T_A is the theoretical minimum of the system noise-temperature T_s which would be found in an *ideal* noisefree receiving system ($T_T = 0$ in equation 1):

$$T_{s\,ideal} = T_A. \tag{2}$$

For a real system with $T_T \neq 0$ the noise performance is best described by the ratio:

$$F_s = \frac{T_s}{T_{S\,ideal}} = \frac{T_A + T_T}{T_A} = 1 + \frac{T_T}{T_A} \tag{3}$$

F_s can be called the operational noise-figure, but F_s is different from other known definitions of noise figures, which use T_o = 300 K instead of T_A and give no information about the system operating under real conditions. $F_s = 1$ is the theoretical minimum and T_T/T_A in equation (3) is the relative contribution of the electronic noise to the system noise. A good noise-performance of the system is found as long as $T_T/T_A < 1$ or $T_T < T_A$ or $F_s < 2$. Then the system is very near to the theoretical noise-minimum and can not effectively further be improved by lowering T_T. Therefore, it is recommended that a system with $F_s < 2$ is called an optimum system.

The noise of a transistor can be described by two equivalent noise sources which are partially correlated. For better understanding of active-antennas it is recommended to introduce these two noise sources as a noise voltage-source u_N and a noise current-source i_N, as in Fig. 7. In Fig. 7 Z_A is the impedance measured at the output terminals 2-2′ of the passive antenna with R_1 between terminals 1-1′. Z_T is the input impedance of the transistor at terminals 3-3′. The noise voltage u_{N3} generated by the two noise sources between terminals 3-3′ is the sum of u_N and the additional noise voltage which is generated by the current i_N flowing through Z_A and Z_T in parallel. The formula for u_{N3} can be very complicated if u_N and i_N are partially correlated. Therefore, transistor noise in active antennas is mostly studied experimentally. It is easily understood that the noise voltage u_{N3}, and the equivalent noise-temperature T_T, depend considerably on the impedance Z_A which the antenna offers to the transistor at terminals 2-2′. Also the signal voltage at terminals 3-3′, which comes from the signal source, depends considerably on the impedance transformation in the passive antenna between terminals 1-1′ and 2-2′. Therefore the signal-to-noise ratio at terminals 3-3′ depends on Z_A in a very complicated way [13, 14]. There is an optimum value, of Z_A, called Z_{Aopt}, which gives best signal-to-noise ratio and lowest transistor noise-temperature T_{Tmin}. This effect is called 'noise match'. If the antenna is lossfree and if the antenna-impedance is transformed into Z_{Aopt} at terminals 2-2′ the receiving system operates with lowest possible system-noise temperature:

$$T_{smin} = T_A + T_{Tmin}. \tag{4}$$

A curve of possible T_{Tmin} for modern transistors is found in Fig. 6. Today for frequencies below 1 GHz $T_{Tmin} < T_A$ and $T_{smin} < T_A$ can be realized. Therefore, for frequencies below 1 GHz active antennas can be designed with optimum noise-performance $F_s < 2$, as defined before. Improved transistors will soon enable us to use active antennas also at microwave frequencies beyond 1 GHz. Our experience with many types of active antennas has shown that noise match in integrated antennas gives an effective improvement of signal-to-noise ratio in real systems, even in mass production.

Noise match is different from power match. Power match would exist when the input network N* transforms the signal in such a way that for a given signal source one gets maximum signal power in Z_T, i.e. maximum signal amplitude at terminals 3-3′ and, consequently, maximum signal output of the active antenna at terminals 6-6′. But this power match is far from giving optimum signal-to-noise ratio. If in Fig. 3a the line section is matched to its characteristic impedance at both ends power match is obtained between signal source and transistor. This matched line section, which is mostly used today, cannot give noise match and, therefore, cannot lead to the optimum signal-to-noise ratio.

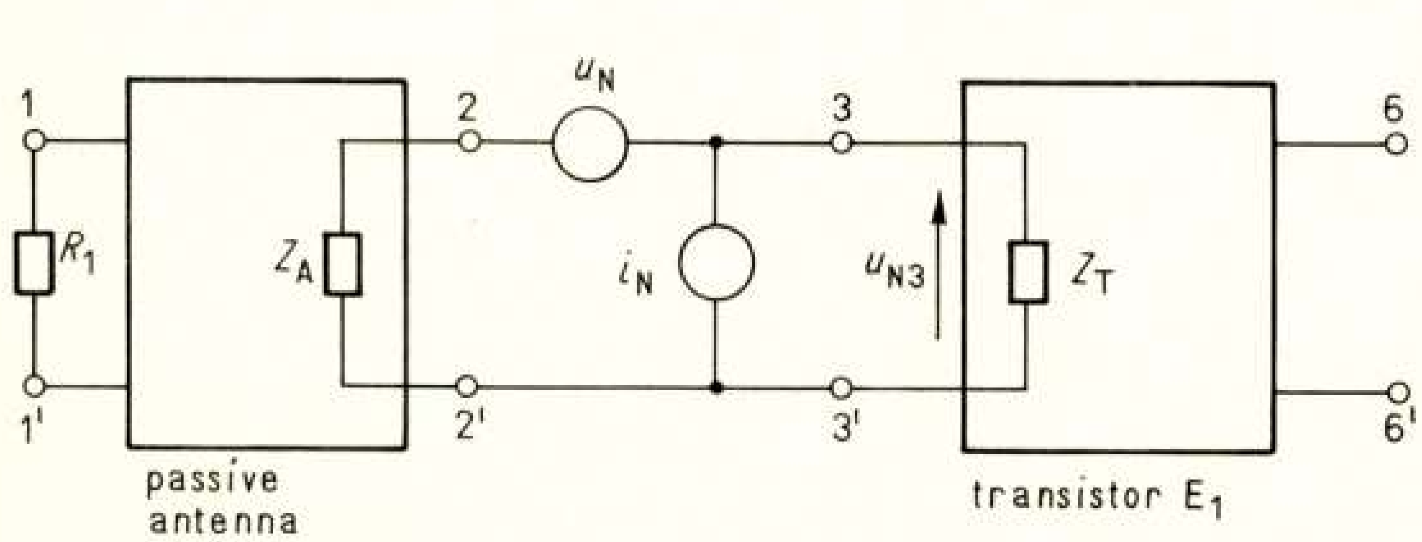

Fig. 7. Equivalent noise sources of the transistor.

3. Low-noise bandpass-antennas

The main problem with noise match is that for all real passive antennas the output impedance Z_A is frequency dependent. So noise match is only possible at a few frequencies. If in Fig. 7 $Z_A = R_A + jX_A$ differs from Z_{Aopt} the transistor noise-temperature T_T is higher than T_{Tmin}. T_T increases with increasing deviation of Z_A from Z_{Aopt}. Circles of constant noise temperature can be drawn in the complex impedance plane of Z_A [14]. Fig. 8 shows such circles of constant noise temperature for

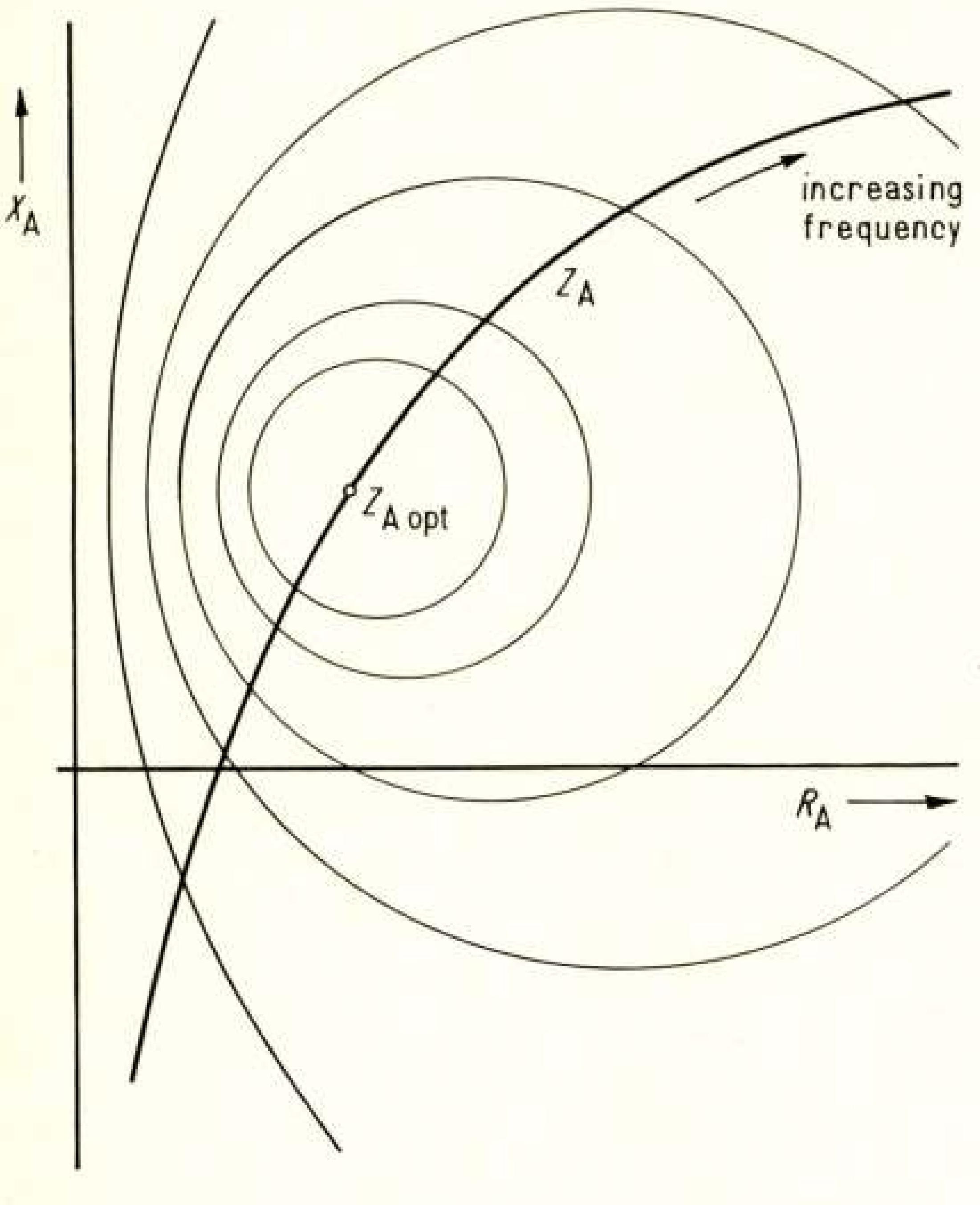

Fig. 8. Circles of constant transistor noise-temperature in the complex impedance plane and the antenna impedance Z_A.

example. These circles have to be measured for the given transistor type at the prescribed frequencies and indicate the noise-temperature T_T for each value of the complex impedance Z_A. One also has to measure Z_A and to draw a curve of Z_A into the circle diagram of Fig. 8 in its dependence on frequency. If this Z_A-curve goes through Z_{Aopt} noise match is obtained for that single frequency f_o, for which $Z_A = Z_{Aopt}$. For all other frequencies $T_T > T_{Tmin}$ and the curve of T_T will depend on frequency like curve 1 of Fig. 9. For a broadband receiving system

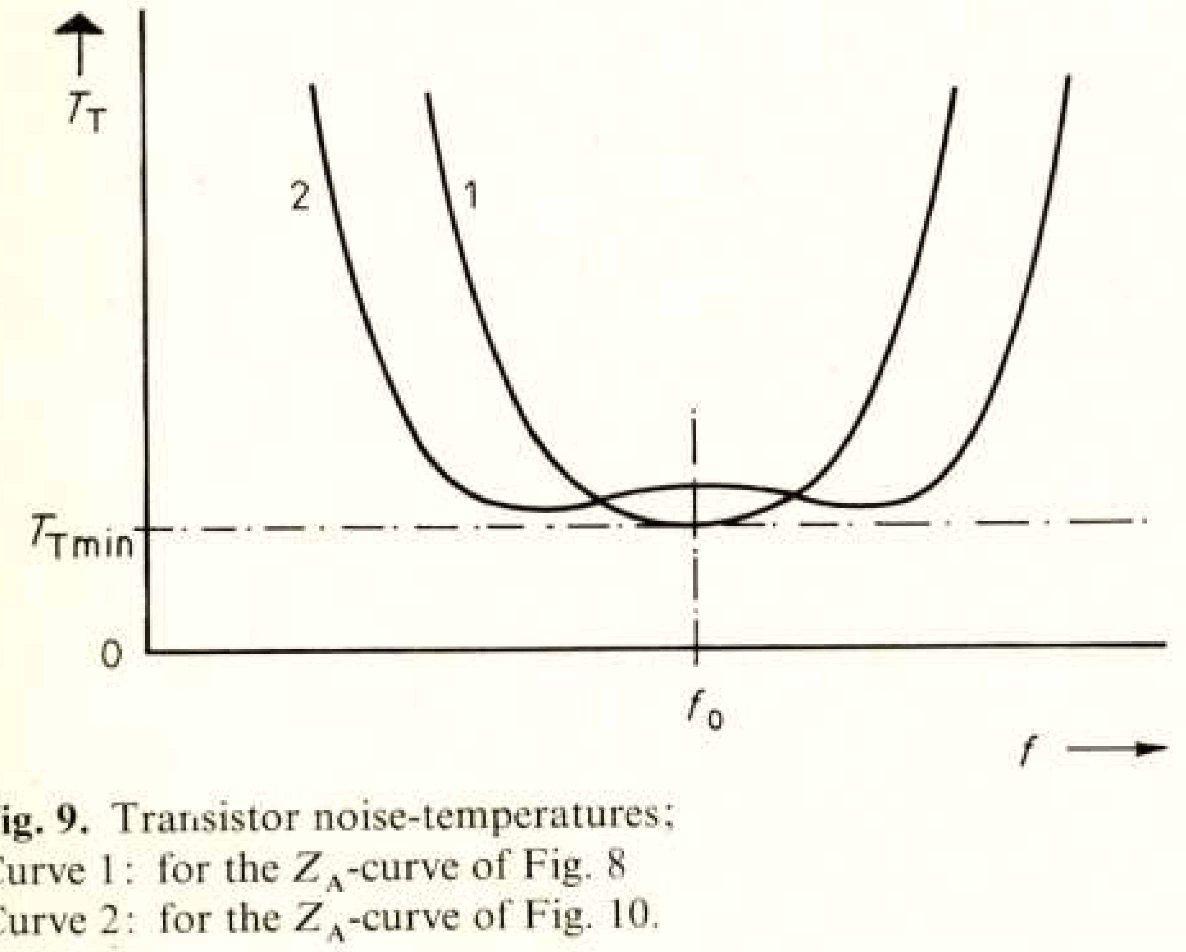

Fig. 9. Transistor noise-temperatures;
Curve 1: for the Z_A-curve of Fig. 8
Curve 2: for the Z_A-curve of Fig. 10.

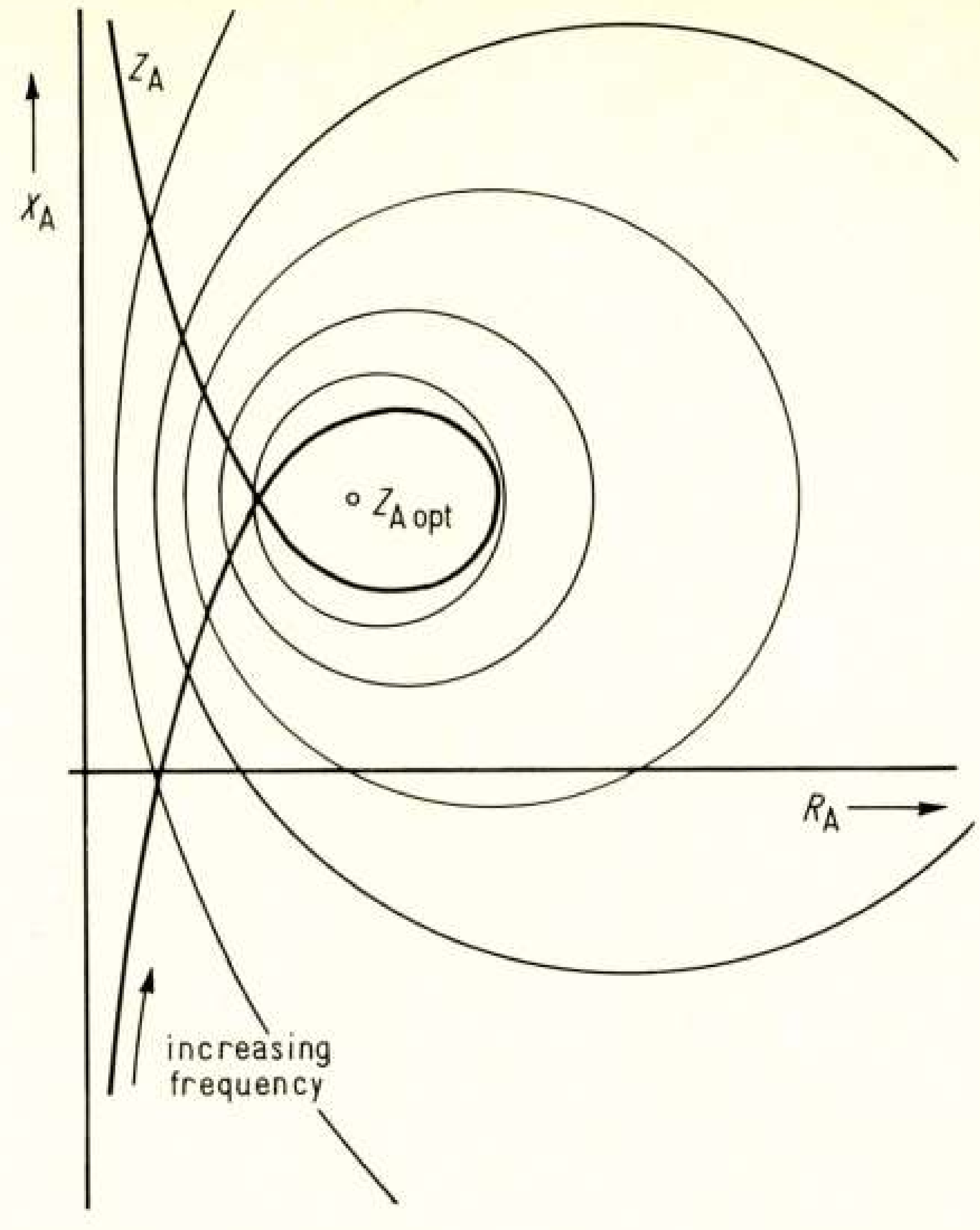

Fig. 10. Antenna impedance Z_A of a bandpass antenna.

we need a low and nearly constant T_T in a broad frequency band. A first step towards satisfying this requirement is a bandpass structure of the integrated antenna, for which with increasing frequency the impedance Z_A forms a loop around Z_{Aopt} in the complex impedance plane as shown in Fig. 10. This gives a noise-temperature T_T in accordance with curve 2 in Fig. 9, which in a broad frequency band is very near to T_{Tmin}. Such an integrated bandpass-antenna is shown in Fig. 4. Antennas of this kind exist commercially for the frequency-modulation radio-band between 88 and 108 MHz [6], for the navigational band between 108 MHz and 156 MHz [8] and for the three television bands [7]. All these antennas are in production and in practical use.

4. Wide-band low-frequency antennas

Noise match, as described before, is only useful at frequencies above 50 MHz. According to Fig. 6 the antenna noise T_A is so high for frequencies below 50 MHz that in equation (4) T_{smin} is nearly identical with T_A, and T_{Tmin} can be neglected. Therefore, at frequencies below 50 MHz the transistor can have higher noise temperatures and noise-mismatch. As long as $T_T < T_A$ the receiving system has optimum noise conditions ($F_s < 2$) as defined before. Therefore, the impedance Z_A can be far from Z_{Aopt}. This fact allows the design of a new type of active antennas for lower frequencies. These are very short integrated monopoles for a very wide frequency band, for example for the radio band from 150 kHz to 15 MHz. This low-frequency antenna can be explained in a simplified form by the help of Fig. 11, which is derived from the general form of Fig. 2. The short monopole is an antenna with an extremely small radiation resistance R_1 and its reactive components are mainly two capacities, a capacity C_{A1} in series to R_1 and a capacity C_{A2} parallel to the output terminals 2-2'. Division of the antenna capacity into these two capacities is the result of a very modern dipole theory and has been verified by experiments. C_{A1} is the receiving capacity in series with the source. C_{A2} is a non-receiving stray capacity between the antenna conductors in the nearfield. An example with high stray capacities of this kind is found with many car antennas which are very near to the metallic car body; especially some modern wire antennas are mentioned here that are installed in the front window. These can show large stray capacities C_{A2} against the conducting frame of the window if the wire is very near to that frame.

At lower frequencies the passive transistor input in Fig. 11 is

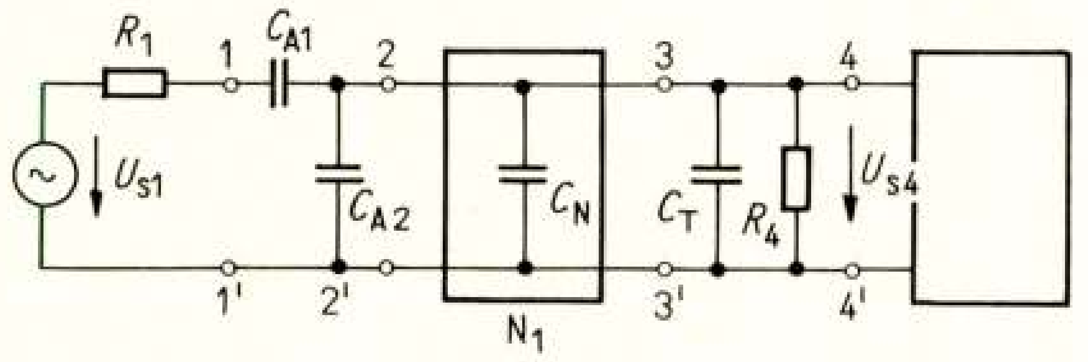

Fig. 11. Input circuitry at low frequencies for a receiving system with short monopole.

a capacity C_T. As in Fig. 2 N_1 is the passive network between the passive antenna and the transistor input. For an antenna with a very broad frequency band no inductances are allowed in N_1 to avoid any resonances. Therefore, no matching is possible between antenna impedance and transistor input because matching of capacitive parts needs inductances. The network N_1 is mainly a shunt capacity C_N due to the connecting leads between antenna and the transistor. The signal voltage U_{s4} which drives the transistor at terminals 4-4′ is smaller than the received signal voltage U_{s1} of the source at terminals 1-1′. The capacities form a voltage divider as shown in Fig. 12 in which:

$$C_P = C_{A2} + C_N + C_T \tag{5}$$

is the sum of all shunt capacities. This voltage divider is nearly frequency-independent under the following conditions: The monopole is sufficiently short and, consequently, R_1 will be so small compared to the reactance $1/\omega C_{A1}$ that it can be neglected; the resistance R_4 of the transistor is so high compared to the reactance $1/\omega\ C_P$ that the effect of R_4 can be neglected. With $R_1 = 0$ and $R_4 = \infty$ we get in Fig. 12:

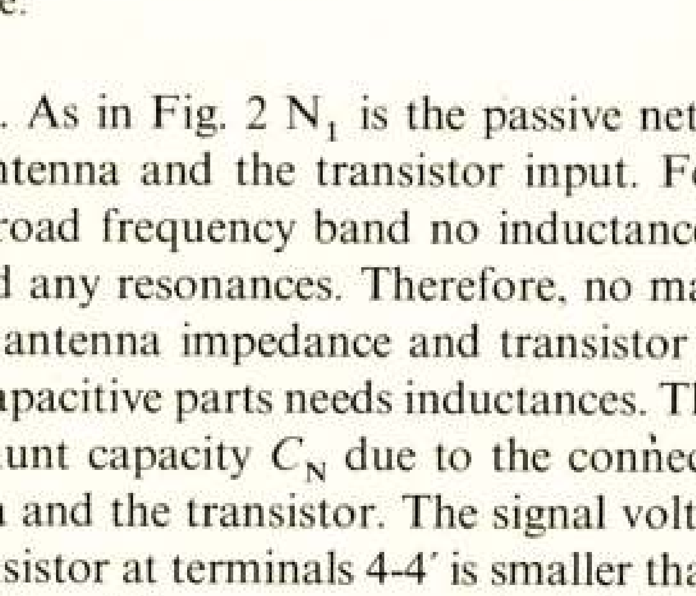

Fig. 12. Equivalent circuit for the configuration of Fig. 11 in which transistor noise-sources are included.

$$U_{s4} = U_{s1} \cdot \frac{C_{A1}}{C_{A1} + C_p} \tag{6}$$

This is a frequency-independent relationship. R_1 and R_4 make the voltage division slightly frequency-dependent and with R_1 and R_4 in the circuitry U_{s4} will be smaller than in equation (6).

As far as the signal-to-noise ratio is concerned one has to make the signal voltage U_{s4} as high as possible to compete with transistor noise. It can easily be seen from equation (6) that C_P must be as small as possible to make U_{s4} as high as possible. So in a first step the signal amplitude U_{s4} is improved by combining the transistor directly with the monopole so that in equation (5) $C_N = 0$. In a second step a transistor type with high R_4 and small C_T is chosen.

In a third step it will be very successful to design and mount the monopole in such a way that the non-receiving capacity C_{A2} is as small as possible. By many experiments we have shown the excellent performance of such wide-band low-frequency antennas. By minimizing C_P either the signal-to-noise ratio of an active monopole of given length can be improved or equal noise performance can be obtained with shorter monopoles. An impressive improvement is obtainable with these minimum-capacity active antennas compared with conventional antennas.

To calculate the signal-to-noise ratio the two noise sources u_N and i_N of Fig. 7 will have to be added to the circuitry of Fig. 12. The full noise theory of this antenna cannot be given in this short article. The main point is, that a transistor with low current-noise source i_N will have to be chosen, because the input circuitry of this low-capacity system has a very high impedance at low frequencies and i_N, flowing through a very high impedance, would give high noise-voltage. For each type of transistor there is a special optimum collector current and an optimum capacity combination which will give the best signal-to-noise ratio for a monopole of given height.

5. Nonlinear effects

As in all receiving systems the linearity of the input transistor introduces a special and very important problem with regard to avoiding cross modulation, intermodulation and frequency conversion. A smallband tuned system has no linearity problems because unwanted frequencies are suppressed by the filter and only the wanted signal arrives at the transistor input. So our tuned active antennas show no nonlinearity effects. But all untuned broadband antennas offer nonlinearity problems because many unwanted signal frequencies enter the input network. This is especially dangerous for mobile antennas on cars and ships which sometimes are neighboured by powerful radio transmitters and then receive strong unwanted signals.

At first we had to develop new methods to measure the nonlinearity of active antennas [16]. The surprising result was that our active antennas show a better linearity than was expected and that the measured nonlinearity was mostly the nonlinearity of the receiver. The higher nonlinearity of the receiver is due to the fact that an active antenna offers more output signal to the receiver input than the conventional passive antenna. Therefore, the measured nonlinearity of active antennas will have to be split up carefully into the nonlinearity of the antenna and the nonlinearity of the following receiver. Fig. 3a shows that for a conventional receiver with passive antenna the electronic element E_1 is the low-noise input of the receiving system and is built into the receiver. In an active antenna E_1 is also a low noise input, but built into the antenna. Combining an active antenna with a conventional receiver would mean that one has a low-noise input-amplifier in the antenna followed by a second low-noise amplifier in the conventional receiver. This is not meaningful because a receiving system needs only one low-noise input-amplifier. So it must be pointed out that the active receiving antennas finally will develop their full possibilities only in combination with a new type of receiver, the input of which is not anymore designed for low noise but for best linearity. We developed such a receiver and so could demonstrate this fact. Nevertheless, the active antennas show also in combination with conventional receivers improved performance compared to conventional receivers alone. More than 100 000 active antennas which are already in operation show this improvement. We only like to express that future development of active antennas will still show further progress if new receivers are developed.

The excellent linearity of active receiving antennas is due to several effects. Nonlinearity decreases very fast with decreasing amplitudes. So at first we try to make the input signal level of all interfering signals as small as possible. In this respect it is an important fact that the active antenna is shorter than the conventional antenna and, by consequence, all received signal amplitudes are smaller. Furthermore, in our active antennas the frequency band of the antenna network is limited to the prescribed bandwidth by filters so that unwanted signals outside the operational band of the receiving system do not arrive at the transistor input. So the number of possibly interfering signals of adequate amplitude is small. In addition, our active antennas have a reverse feedback to improve linearity. We mostly use a combination of two transistors instead of one and so get simultaneously low-noise performance and linearity. Such an optimum transistor-combination is only found by the use of a computer program because too many parameters are involved.

A next step in active antenna development is the directional antenna. Here two ways can be followed: Combine one single active antenna with known passive means, for example Yagi structures, lenses, horns and paraboloids. This is a very successful method and is already in practical use. The second way is to combine two or more active dipoles in a directive system. Our preliminary experiments show that completely new antenna forms can be designed with interesting performance. By changing bias we can get steerable patterns.

In these days in all parts of the world research and development on active antennas is going on and production has already started. The number of papers on active antennas presented at conferences and in journals is increasing. So we think that the active antenna has a good future.

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Korte technische berichten

Magneetband-viewer voor het controleren van opgenomen sporen

Een vaak toegepaste wijze voor het zichtbaar maken van op magneetband vastgelegde digitale en analoge signalen bestaat hierin, dat men op de band een dun laagje ijzerhoudend poeder aanbrengt of een geschikte ijzerhoudende vloeistoffilm. Na de verrichte controle moet de aangebrachte laag dan weer zorgvuldig worden verwijderd. Dit is meestal een tijdrovende bezigheid, vooral wanneer de controle tot een aantal bandgedeelten wordt uitgestrekt.

Door Minnesota (Nederland) N.V. wordt een viewer voor magneetband op de markt gebracht, die de te controleren band volkomen schoon laat. Qua omvang en vorm is deze viewer vergelijkbaar met een voorzetlens of kleurenfilter voor een camera. Hij heeft een diameter van $4\frac{1}{2}$ cm en bestaat uit twee plastic schijfjes, waartussen zich een vloeistof bevindt die deeltjes ijzeroxyde bevat. Het geheel is 4 mm dik. Wordt dit handige instrumentje op een magneetband gelegd, dan rangschikt het

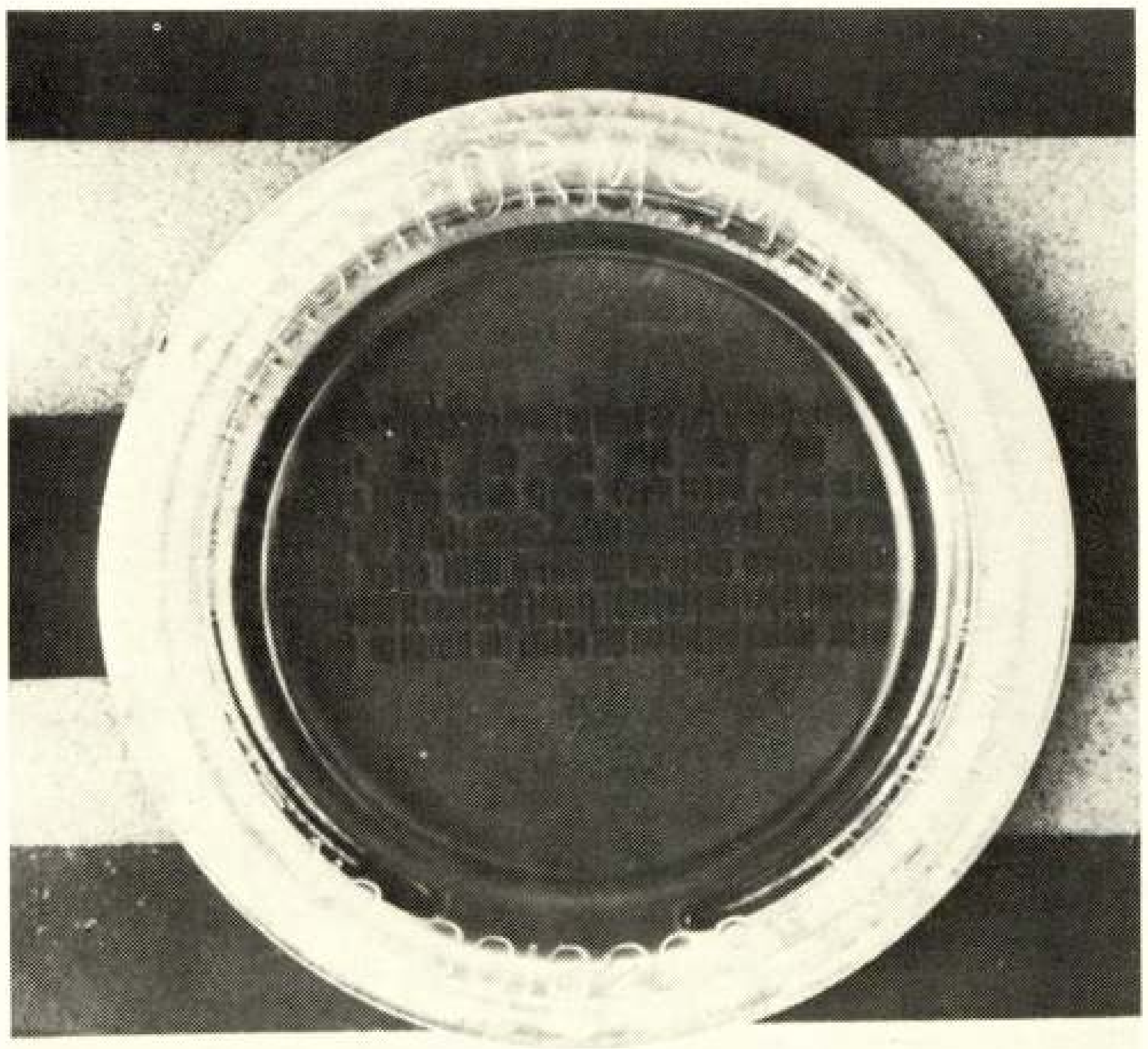

Fig. 1. Tape viewer, die het magnetisch spoor zichtbaar maakt.

ijzeroxyde in de vloeistof zich in een patroon, bepaald door de magnetisatie op de band, waarbij de deeltjes als donkere concentraties zichtbaar worden tegen een lichte achtergrond (fig. 1). Na gebruik kan men de viewer door licht wrijven over de achterzijde weer egaal van tint maken, geschikt voor een volgende controle.

Met deze viewer, door Minnesota aangeduid als 'bx-1022 Plastiform Magnetic Viewer', kan men de uitlijning van magneetkoppen, plaatsing van de sporen, definitie van de pulsen, blokafstanden, enz. snel en gemakkelijk controleren. Bij het opsporen van storingen in recorders kan men uit het feit, of er wel of geen signaal in de viewer zichtbaar wordt, afleiden of de fout in de opname dan wel in de weergave schuilt.

*

3M-Minnesota (Nederland) N.V., Leiden.

Varia

International Journal of Circuit Theory and Applications

During the past number of years the field of circuit theory has expanded considerably and the areas in which it is being applied are becoming ever more numerous. In view of this situation it has been felt worthwhile to establish the '*International Journal of Circuit Theory and Applications*'. The Journal aims to bring together papers on a wide variety of problems encompassing both advances in circuit theory and design and in the uses of circuit theory in other fields. The essential feature of the material to be published is that the *ideas* of circuit theory (taken, of course, to include its implementation by computer) should have played a prominent part in the solution or elucidation of the problem.

An Editorial Board, drawn from a wide variety of countries, will preside over the Journal in order to maintain contact with the subject area on a worldwide basis, and will play an essential part in the refereeing of contributions. It is hoped that the new Journal will provide in some respects a unique contribution because its coverage is worldwide and it specifically includes many of the applications of circuit theory.

The scope of the Journal is intended to cover all aspects of the theory and design of analogue and digital circuits together with the application of the ideas of circuit theory to a wide variety of problems. Examples of the areas covered are: Fundamental circuit theory, circuit modelling of devices, synthesis and design of filters and active circuits, distributed circuits, applications to microwave components and systems, solid state devices and biological systems. Contributions to computer analysis, design and simulation are welcome.

Contributions may consist of, (a) Regular Papers, (b) Short Communications, (c) Letters to the Editor. Short Communications are refereed to the same standards as regular papers but may describe incomplete work or be concerned with a more narrowly defined problem. Letters to the Editor (each limited to 500 words) are subject only to brief refereeing by the Editors and generally achieve more rapid publication. A letter to the Editor would typically comment on published results, pose some new problem, draw attention to some application or otherwise be of technical interest. Contributions may be of a research or tutorial nature.

The Journal will be published by John Wiley & Sons Ltd as a quarterly, beginning in January 1973. Intending contributors may obtain further details from the Editors or any member of the Editorial Board. A prospectus will be available in due course from the publishers, and copies may be obtained from John Wiley and Sons Limited, Baffins Lane, Chichester, Sussex, England.

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Prof. dr. ir. W. Th. Bähler 80 jaar

Op 16 mei jl. vierde de oud-hoogleraar prof. dr. ir. W. Th. Bähler in Doorn, waar hij sinds zijn afscheid van de Technische Hogeschool te Delft woonachtig is, zijn 80e verjaardag. Prof. Bähler geniet evenals zijn echtgenote gelukkig nog een goede gezondheid.

Bij de vele gelukwensen, die in Doorn binnenkwamen, verdient wellicht één bijzondere vermelding. Ter gelegenheid van de verjaardag van prof. Bähler werd namelijk besloten, een opdracht op te nemen in de definitieve uitgave van het eindverslag van het telecommunicatiecolloquium, dat in 1970-1971 aan de Technische Hogeschool te Delft werd gehouden. Deze opdracht luidt:

'*Opgedragen aan de grondlegger van het technisch wetenschappelijk onderwijs in de telecommunicatietechniek in ons land, ter gelegenheid van zijn 80e verjaardag, 16 mei 1972*'.

Het colloquium had als onderwerp 'Het lokale telecommunicatienet van straks'. De definitieve uitgave van het eindverslag zal worden verzorgd door de Stichting Toekomstbeeld der Techniek, ressorterende onder het Koninklijk Instituut van Ingenieurs.

*

T.H. Delft.

Prof. dr. ir. J. P. M. Schalkwijk gewoon hoogleraar aan de T.H. Eindhoven

Prof. dr. ir. J. P. M. Schalkwijk is bij Koninklijk Besluit nr. 25 van 30 maart 1972 benoemd tot gewoon hoogleraar in de telecommunicatie (Afdeling der Elektrotechniek) aan de T.H. Eindhoven.

Prof. Schalkwijk werd op 1 november 1936 te Rijswijk (Z.H.) geboren. In 1959 verkreeg hij aan de T.H. Delft bij de Afdeling Elektrotechniek het ingenieursdiploma. Van 1963 tot 1965 studeerde hij Electrical Engineering aan de Stanford University, welke studie hij afsloot met het behalen van de Ph.D.-graad.

Sedert 1968 is prof. Schalkwijk Assistant Professor of Information and Computer Science aan de University of California in San Diego.

Van zijn hand verscheen reeds een groot aantal publikaties in nationale en internationale tijdschriften en een aantal congresverslagen.

*

Ir. J. H. Geels lector aan de K.M.A.

Ir. J. H. Geels is bij Koninklijk Besluit met ingang van 1 september 1972 benoemd tot gewoon lector in de elektrotechniek aan de K.M.A. te Breda.

*

Toekenning Vederprijs voor 1971

Aan de heer F. L. J. Sangster te Aalst (N.Br.) en aan ir. H. L. Bakker te Hilversum, werd de Vederprijs met gouden medaille toegekend voor het jaar 1971. De feestelijke uitreiking door mevrouw C. E. van Hoboken-Veder vond plaats tijdens een gecombineerde werkvergadering van het NERG en de Sectie voor Telecommunicatietechniek van het Koninklijk Instituut van Ingenieurs op donderdag 27 april 1972, bij de T.H. te Delft, in het het gebouw van de Afdeling Elektrotechniek.

Fig. 1. De heer Bakker (links) en de heer Sangster (rechts) bezien met belangstelling de door hen ontvangen gouden Vedermedailles.

De prijs werd aan de heer Sangster toegekend in verband met een door hem uitgevonden en gerealiseerde categorie van schakelingen, bekend onder de naam van 'emmertjesgeheugen'. Met behulp van dergelijke schakelingen kan met een minimum aan elementen een analoog schuifregister worden gerealiseerd. Dergelijke schuifregisters, in het bijzonder in de vorm van geïntegreerde circuits, kunnen voor een grote verscheidenheid van vertragingstijden worden ontworpen.

De heer Sangster is wetenschappelijk medewerker bij het Philips' Natuurkundig Laboratorium.

Aan ir. Bakker werd de prijs toegekend in verband met door hem ontworpen lijnversterkers voor de versterking van frequentiemultiplex signalen in lange-afstand kabelsystemen. De door hem ontworpen lijnversterkers kenmerken zich door het bijzonder lage ruisniveau van de ingangstrappen, terwijl in de daarbij behorende uitgangstrappen een hoog rendement behaald wordt. Bovendien werd door hem bereikt, dat door de aandacht voor het gehele kabelsysteem een zeer goede temperatuurcompensatie werd verkregen voor de, met de temperatuur van de bodem in eigenschappen veranderende, kabeladerparen.

Hierdoor kunnen coaxiale transmissiestelsels worden gerealiseerd, waarvan de demping over een frequentieband van 12 MHz een ongekend goede lange-duur stabiliteit vertoont.

De heer Bakker is wetenschappelijk medewerker bij Philips' Telecommunicatie Industrie.

De resultaten, die door elk der beide prijswinnaars werden verkregen, hebben internationale belangstelling gewekt.

*

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