

MICROWAVE RADIOMETRY OF VEGETATION CANOPIES

ADVANCES IN GLOBAL CHANGE RESEARCH

VOLUME 24

Editor-in-Chief

Martin Beniston, *Department of Geosciences, University of Fribourg, Switzerland*

Editorial Advisory Board

- B. Allen-Diaz, *Department ESPM-Ecosystem Sciences, University of California, Berkeley, CA, U.S.A.*
- R.S. Bradley, *Department of Geosciences, University of Massachusetts, Amherst, MA, U.S.A.*
- W. Cramer, *Department of Global Change and Natural Systems, Potsdam Institute for Climate Impact Research, Potsdam, Germany.*
- H.F. Diaz, *Climate Diagnostics Center, Oceanic and Atmospheric Research, NOAA, Boulder, CO, U.S.A.*
- S. Erkman, *Institute for Communication and Analysis of Science and Technology – ICAST, Geneva, Switzerland.*
- R. García Herrera, *Facultad de Físicas, Universidad Complutense, Madrid, Spain*
- M. Lal, *Centre for Atmospheric Sciences, Indian Institute of Technology, New Delhi, India.*
- U. Luterbacher, *The Graduate Institute of International Studies, University of Geneva, Geneva, Switzerland.*
- I. Noble, *CRC for Greenhouse Accounting and Research School of Biological Sciences, Australian National University, Canberra, Australia.*
- L. Tessier, *Institut Méditerranéen d'Ecologie et Paléocologie, Marseille, France.*
- F. Toth, *International Institute for Applied Systems Analysis, Laxenburg, Austria.*
- M.M. Verstraete, *Institute for Environment and Sustainability, EC Joint Research Centre, Ispra (VA), Italy.*

The titles published in this series are listed at the end of this volume.

MICROWAVE RADIOMETRY OF VEGETATION CANOPIES

by

Alexander A. Chukhlantsev

Russian Academy of Sciences, Moscow, Russia

 Springer

A C.I.P. Catalogue record for this book is available from the Library of Congress.

ISBN-10 1-4020-4681-2 (HB)
ISBN-13 978-1-4020-4681-0 (HB)
ISBN-10 1-4020-4682-0 (e-book)
ISBN-13 978-1-4020-4682-7 (e-book)

Published by Springer,
P.O. Box 17, 3300 AA Dordrecht, The Netherlands.

www.springer.com

Printed on acid-free paper

All Rights Reserved

© 2006 Springer

No part of this work may be reproduced, stored in a retrieval system, or transmitted in any form or by any means, electronic, mechanical, photocopying, microfilming, recording or otherwise, without written permission from the Publisher, with the exception of any material supplied specifically for the purpose of being entered and executed on a computer system, for exclusive use by the purchaser of the work.

Printed in the Netherlands.

TABLE OF CONTENTS

Foreword	ix
Preface	xi
Chapter 1. BASICS OF MICROWAVE RADIOMETRY	1
1.1. Theory of Thermal Radiation	1
1.2. Microwave Radiometers and Systems	12
1.3. Radiometric Measurements	18
Chapter 2. PHYSICAL AND ELECTRICAL PROPERTIES OF SOILS AND VEGETATION	21
2.1. Physical and Dielectric Properties of Soils	21
2.2. Biometrical Features and Electrical Properties of Vegetation	36
Chapter 3. MICROWAVE EMISSION FROM BARE SOILS	53
3.1. Microwave Emission Models of Bare Soils	53
3.2. Experimental Research on Microwave Emission from Bare Soils.....	69
Chapter 4. THEORY OF MICROWAVE PROPAGATION THROUGH VEGETATION MEDIA	75
4.1. General Approach to the Description of Electromagnetic wave Propagation in Vegetation	75
4.2. The Model of Vegetation as a Continuous Medium	78
4.3. The Model of Vegetation as a Collection of Scatterers (Discrete Model)	91
4.4. Extinction and Scattering of Electromagnetic Waves by Plant Elements	97

4.5. Microwave Propagation Through a Vegetation Layer. Relation of Electrodynamical Parameters to Biometric Features of Vegetation	110
Chapter 5. EXPERIMENTAL STUDIES OF MICROWAVE PROPAGATION IN VEGETATION CANOPIES	119
5.1. Methods of Experimental Research	119
5.2. Experimental Results	129
Chapter 6. MODELING OF MICROWAVE EMISSION FROM VEGETATION CANOPIES	147
6.1. General Approach to the Modeling of Radiation Parameters for Vegetated Soils	147
6.2. The Emissivity of a Vegetation Canopy	154
6.3. Modeling Microwave Emission from Forests	166
6.4. Polarization Properties of Microwave Emission from Vegetation Canopies	169
6.5. Spatial Variations of Microwave Emission from the Earth's Surface	171
6.6. Global Simulation of Microwave Emission from Land	173
Chapter 7. EXPERIMENTAL RESEARCH ON MICROWAVE EMISSION FROM VEGETATION CANOPIES	177
7.1. Research on Microwave Emission from Vegetated Fields ...	177
7.2. Research on Microwave Emission from Forests	197
7.3. Statistical Properties of Microwave Emission from Vegetation Canopies	204
Chapter 8. VEGETATION EFFECT IN MICROWAVE REMOTE SENSING	207
8.1. Accounting for Vegetation Effect in Microwave Radiometry of a Surface	207
8.2. Vegetation Biomass Retrieval from Microwave Radiometric Measurements.....	217
8.3. Soil Moisture and Vegetation Biomass Retrieval from Multi-Configuration Microwave Radiometric Measurements.....	230
8.4. Vegetation Effect in Active Microwave Remote Sensing ...	236

Chapter 9. MICROWAVE RADIOMETRY OF VEGETATION CANOPIES IN CONTEXT OF GLOBAL CHANGE RESEARCH	241
9.1. Global Climate Problems and the Carbon Cycle	241
9.2. Assimilation of Remote Sensing Data into Global Carbon Cycle Models	248
References	257
Index	281

FOREWORD

Research into microwave radiation from the Earth's surface in the presence of vegetation canopies, as well as the development of algorithms for retrieval of soil and vegetation parameters from microwave radiometric measurements, have been actively conducted for the last thirty years by many scientific groups and organizations all over the world. A complete bibliography of works on this problem would encompass hundreds of titles. The capability of the microwave radiometric method to determine soil moisture and vegetation biometric indices was revealed a quarter of a century ago by the author and many of his colleagues. In spite of the fact that the fundamentals and the basic physics of the microwave radiometry of soils and vegetation covers have been well developed, interest in the problem has not decreased but indeed has grown significantly in the last decade. This phenomenon has several reasons. The first one is the importance of these objects themselves in the remote ecological monitoring of land surface. In fact, soil moisture and vegetation covers play a key role in the hydrological cycle and in water and energy transfer on the border of land surface and atmosphere through evaporation and transpiration. The second reason is increased technical potentialities of microwave radiometric devices by being installed on spacecraft. In the modern design of space microwave radiometric systems, a high spatial resolution is achieved by using multi-beam antennas, synthetic aperture antennas, and big antennas with electronic and mechanical scanning that allows obtaining radio images of the Earth's surface. Accomplishment of large international projects that include global monitoring of the hydrological state of land surface (EOS Aqua, SMOS, Hydros, and others) shows that microwave radiometry of soil and vegetation more and more has become an instrument of practical application and operational use. In this respect, a systematic account of questions concerning the microwave radiometry of the Earth's surface in the presence of vegetation canopies seems to be useful and is the main objective of this book. The fundamental three-volume book by Ulaby, Moore, and Fung (1981-1986.), in which these questions were partially considered, was published a long ago. For a time after its publication, many papers on the

problem appeared. A large number of papers and books were published in Russian and are not widely available for western researchers. To systemize these publications is also an objective of this book. The other objective of this monograph is to present young scientists, MS and PhD students who are involved or intending to be involved with microwave radiometric programs, with a text-book for studying the problem (and microwave radiometry in general) as a whole. Therefore, the book includes both general questions about microwave radiometry and particular practical questions of microwave radiometric measurements, design of experiments, etc.

The book is mainly based on works completed in the Institute of Radioengineering and Electronics of the Russian Academy of Sciences by the author and his colleagues during the last thirty years. Results by other researchers and groups are also reviewed, summarized, and analyzed. The author tried to review works of all researchers involved with the considered problem. But the number of works on the topic is really big, and he is sorry if someone's contribution is not mentioned in the book. The author will be glad to receive remarks and suggestions concerning the content of the book.

PREFACE

During the past twenty to thirty years, ferment in globalization processes in all sphere's of human activity has revealed numerous problems that arise from the interaction of society with nature. Among these problems, the global carbon cycle has acquired a special significance because of its now well-known greenhouse effect. To solve this and many similar problems, it is necessary to develop new concepts and approaches to our analyses of global environmental change, and in particular to select priorities in observing and assessing the existing state of natural systems. For example, at the present time, different global change models are being developed to assist in predicting climatic change. The functioning of these models and the reliability of estimates provided by them require permanent monitoring of the natural systems state. This monitoring allows us to detect real changes in environmental conditions and to compare predicted and observed trends in these changes. In addition, the observed data on the state of environmental components may be used as both model input parameters and feedback for model corrections.

Some ideas of permanent monitoring of environmental change have been realized with Earth Observing Systems that include space-borne and ground-based units to collect data on different parameters of the Earth's surface and atmosphere. Space-borne sensors play a key role in such monitoring, since they can rapidly collect a large amount of data over extensive territories. At the same time, ground-based data are important in validating and verifying the satellite information and filling possible gaps in remote observations.

The space-borne means of Earth observation that have been developed during the last decades include practically all significant functions of the nature-society system. Remote sensing data are the main source of operative information for the systems of control of global ecological, biogeochemical, hydro-physical, epidemiological, geophysical, and even demographic situations on the Earth. Remote sensing of the Earth's surface and atmosphere from space is based on receiving emitted or scattered electromagnetic radiation in different ranges of the electromagnetic wave

spectrum. Operation of space-borne remote sensing means is affected by forces in the atmosphere. Particularly, optical sensors are not efficient in the presence of clouds. They also can not be used at night, since they can only detect solar radiation scattered by the Earth's surface. Infrared sensors are also influenced by the atmosphere. On the other hand, microwave remote sensors operate at decimeter, centimeter, and millimeter wavelengths. The atmosphere is practically transparent at decimeter and centimeter wavelengths, and microwave observations can be performed even in the presence of heavy cloudiness.

Microwave remote sensing methods are divided into active (radar) and passive (radiometric) ones. Microwave radiometric sensors measure the parameters of thermal electromagnetic radiation from the Earth's surface that provides their twenty-four-hour performance. Different remote sensing means are sensitive to various atmosphere and surface parameters. It appeared that microwave radiometry as applied to observations of land surface is an appropriate instrument for monitoring surface soil moisture and temperature. Soil moisture plays a crucial role in hydrology, agronomy, and meteorology. It governs the redistribution of precipitation between infiltration and runoff, it affects the development of crops through its dominance on regulating water-uptake by the plants, and it manages the partitioning of energy and water through evaporation and transpiration at the lower boundary of the atmosphere. Soil moisture is thus a key variable in the hydrological cycle. Monitoring soil moisture status on regional and global scales is of primary importance for understanding and protecting the environment, as well as for natural resources management that has been emphasized by the World Climate Research Program (1995). That is why two space missions have been planned, that are directly intended for soil moisture observations at a global scale by means of microwave radiometry.

The European Space Agency has selected the Soil Moisture and Ocean Salinity (SMOS) mission for implementation as the second mission in the line of Earth Explorer Opportunity mission (Silvestrin *et al.*, 2001; Kerr *et al.*, 2001). The goal of the SMOS mission is to observe two key variables, namely soil moisture over land and ocean surface salinity, by means of *L*-band (a wavelength of 21 cm) microwave imaginary radiometry. SMOS will also provide information on root zone soil moisture, vegetation, and biomass and contribute to research on the cryosphere. It is expected that the knowledge of global distribution of soil moisture and ocean salinity at adequate spatial and temporal sampling will significantly enhance weather forecasting, climate and extreme event predictions.

In the USA, the NASA Hydrosphere State (Hydros) mission was selected by NASA in 2002 as an Earth System Science Pathfinder mission for further development and is currently scheduled for launch in 2010

(Entekhabi *et al.*, 2004). Hydros is designed to provide global maps of the Earth's soil moisture and freeze/thaw state every 2-3 days for weather and climate prediction, water, energy, and carbon cycle studies, and natural hazards monitoring. Hydros utilizes a unique active and passive L-band microwave concept to simultaneously measure microwave emission and backscatter from the surface across a wide spatial swath. The key derived products are soil moisture at 40-km resolution for hydro-climatology obtained from the radiometer measurements, soil moisture at 10-km resolution for hydrometeorology obtained by combining the radar and radiometer measurements in a joint retrieval algorithm, and freeze/thaw state at 3-km resolution for terrestrial carbon flux dynamics studies obtained from radar measurements.

The presence of a vegetation cover on the Earth's surface affects the microwave emission from the surface in two ways. First of all, the vegetation cover screens the microwave emission from the soil surface. Secondly, the microwave emission from the vegetation layer itself is added to the emission from the soil. Vegetation effect depends on the vegetation type and vegetation biometric features as well as on the measuring configuration (frequency, polarization, observation angle). Assessment of the vegetation impact on the microwave radiometry of surface soil moisture, and examination of feasibilities of biometric parameters retrieval from microwave radiometric measurements are the main objectives of the present work.

The book contains nine chapters. In Chapter 1 introductory knowledge on the basics of microwave radiometry is given. Numerous books are available now, where the principles and fundamentals of microwave radiometry are developed and stated (e.g., Rytov, 1953; Levin and Rytov, 1967; Basharinov *et al.*, 1968; Basharinov *et al.*, 1974; Rytov *et al.*, 1978; Bogorodskii *et al.*, 1977; Bogorodskii *et al.*, 1981; Ulaby *et al.*, 1981, 1982, 1986; Tsang *et al.*, 1985; Sharkov, 2003; Armand and Polyakov, 2005), and, therefore, the microwave radiometry basics are given in brief. The material presented in Chapter 1 is taken from publications mentioned above.

Physical and microwave dielectric properties of vegetation and soil are discussed in Chapter 2. Several dielectric models of vegetation matter and soil are examined. It is shown that in the microwave band the dielectric properties of soils and vegetation material are mainly determined by their water content.

In Chapter 3, theoretical models and experimental data on the microwave emission from bare soils are presented. It is shown that the main factors that determine the intensity of microwave emission from bare soils are the moisture content of soil top layer and soil surface roughness. Other

factors, such as soil type (texture), soil bulk density, etc., have a minor effect on the soil microwave emission.

The theory of microwave propagation in vegetation canopies is developed in Chapter 4. The relation between microwave propagation characteristics and vegetation biometric parameters are established. It is shown, particularly, the extinction rate of the coherent electromagnetic wave in a vegetation medium is proportional to the vegetation water content in unit volume, whereas the optical depth of a vegetation layer is mainly determined by the vegetation water content per unit area.

Chapter 5 presents a review of experimental research on microwave propagation in vegetation canopies. Different experimental techniques are discussed. Available experimental data on microwave attenuation by vegetation canopies are compared with model simulations.

Theory of microwave emission from vegetation canopies is developed in Chapter 6. A simple emission three-component radiation model (the $\tau-\omega$ model) is propounded. It is shown that the microwave emission of a vegetated soil is determined mainly by soil moisture and roughness, temperature of soil and vegetation, and vegetation water content.

An overview of experimental research on microwave emission from vegetation canopies is presented in Chapter 7.

Vegetation effect in microwave remote sensing of terrains is considered in Chapter 8. Vegetation screening in microwave remote sensing of soil moisture is discussed. Retrieval of soil moisture and vegetation water content from microwave radiometric measurements is examined.

In Chapter 9, a possibility of assimilation of microwave radiometric remote sensing data into global carbon cycle models is taken up.

Chapter 1

BASICS OF MICROWAVE RADIOMETRY

1.1. THEORY OF THERMAL RADIATION

1.1.1. Essential Definitions and Equations of an Electromagnetic Field

A full-blown theory of a unified electromagnetic field was developed by Maxwell. This theory is based on a generalization of known laws, established experimentally, and is a phenomenological macroscopic theory. It does not consider the microscopic processes in a medium in the presence of an electromagnetic field. Electric and magnetic properties of a specific medium are described by three macroscopic quantities, i.e., its relative dielectric permittivity ε , its relative magnetic permeability μ , and its conductivity σ . It is assumed that these parameters are known from experiments. Maxwell's theory solves the major task of electrodynamics, i.e., to determine the electromagnetic fields for a given system of electric currents and charges. The full system of *Maxwell's equations* is given by

$$\operatorname{rot}\vec{E} = -\frac{\partial\vec{B}}{\partial t}, \quad (1.1)$$

$$\operatorname{div}\vec{D} = \rho_q, \quad (1.2)$$

$$\operatorname{rot}\vec{H} = \vec{j} + \frac{\partial\vec{D}}{\partial t}, \quad (1.3)$$

$$\text{div}\vec{B} = 0 \quad (1.4)$$

where \vec{E} is the electric field strength, \vec{B} is the magnetic flux density, ρ_q is the charge density, and \vec{j}_e is the current density. The system is supplemented with the *material equations* that, for isotropic, non-ferroelectric, and non-ferromagnetic media, are written as

$$\vec{D} = \epsilon\epsilon_0\vec{E}, \quad (1.5)$$

$$\vec{B} = \mu\mu_0\vec{H}, \quad (1.6)$$

$$\vec{j}_e = \sigma\vec{E} \quad (1.7)$$

where ϵ_0 and μ_0 are the electric and magnetic constants, respectively.

Maxwell's theory permits the existence of electromagnetic fields in a space even if currents and charges are absent there. These fields are called electromagnetic waves that propagate through the space. In this free space, the propagation of an electromagnetic field is described by the following wave equations, which can be obtained from the above equations of Maxwell's theory:

$$\nabla^2\vec{E} - \epsilon\epsilon_0\mu\mu_0\frac{\partial^2\vec{E}}{\partial t^2} = 0, \quad (1.8)$$

$$\nabla^2\vec{H} - \epsilon\epsilon_0\mu\mu_0\frac{\partial^2\vec{H}}{\partial t^2} = 0. \quad (1.9)$$

Electromagnetic waves are transverse waves and propagate with the phase velocity $v = c/\sqrt{\epsilon\mu}$ where $c = 1/\sqrt{\epsilon_0\mu_0}$ is the velocity of electromagnetic waves in a vacuum. The vectors \vec{E} and \vec{H} are mutually perpendicular and both vectors are perpendicular to the velocity vector $\vec{v} = v[\vec{E} \times \vec{H}]/EH$. The magnitudes of vectors \vec{E} and \vec{H} in a wave are linked as

$$H = \sqrt{\epsilon\epsilon_0/(\mu\mu_0)} E. \quad (1.10)$$

A sinusoidal electromagnetic wave is called a monochromatic wave. If the time dependence of a field is chosen in the form $E \sim \exp\{-j\omega t\}$ where $\omega = 2\pi f$, and f is the frequency, the wave equation takes the form

$$\nabla^2 \vec{E} - \epsilon\epsilon_0\mu\mu_0\omega^2\vec{E} = 0. \quad (1.11)$$

If the wave field depends on only one coordinate, the wave is called a plane wave. In other words, the dependence of a plane wave field on the coordinates is described by the multiplier $\exp\{j\vec{k}\vec{r}\}$ where \vec{k} is the wave vector ($k = \omega/v$). For example, the electric field of the plane monochromatic wave propagating in z direction is found from (1.11) and is given by

$$E_x = E_{x0} \exp\{-j\omega t + k_z z\}, \quad (1.12)$$

$$E_y = E_{y0} \exp\{-j\omega t + k_z z + \varphi\} \quad (1.13)$$

where φ is the phase difference of E_x and E_y oscillations. With an arbitrary value of φ , the plane wave is *elliptically* polarized. This means that in every point of a wave field the end of vector \vec{E} describes an ellipse lying in the x - y plane. If $E_{x0} = E_{y0}$ and $\varphi = \pm(2m+1)\pi/2$, ($m = 0, 1, 2, \dots$), the ellipse transforms into a circle, and the wave is *circular* polarized. If $\varphi = \pm m\pi$, ($m = 0, 1, 2, \dots$), the ellipse turns into a line, and the wave is *linear* polarized. The plane passing through vector \vec{E} and wave vector \vec{k} is called the *polarization plane* of a linear polarized wave. If a linear polarized wave is incident upon a half-space at an incidence angle (the angle between \vec{k} and the perpendicular to the surface of half-space) and the polarization plane is perpendicular to the surface, the wave is considered as *vertically* (v) polarized. If vector \vec{E} of the incident wave is parallel to the surface, the wave is called *horizontally* (h) polarized.

The *volume density of electromagnetic field energy* is defined as the sum of volume densities of electric field energy and magnetic field energy. The volume density of electromagnetic field energy is given by

$$w_e = \frac{\epsilon\epsilon_0 E^2}{2} + \frac{\mu\mu_0 H^2}{2}. \quad (1.14)$$

The volume density of electromagnetic wave energy is written as (taking into account equation (1.10))

$$w_e = \epsilon\epsilon_0 E^2 = EH/v. \quad (1.15)$$

A moving electromagnetic wave transfers electromagnetic energy. The velocity of energy transfer is described by the Poynting's vector that is given by

$$\vec{\Pi} = w_e \vec{v} = [\vec{E} \times \vec{H}]. \quad (1.16)$$

This vector determines the density of electromagnetic energy flux, i.e., it shows what electromagnetic energy passes through a unit area in unit time.

The *intensity* of a moving monochromatic wave is defined as the mean over the period value of Poynting's vector:

$$I = \left\langle \vec{\Pi} \right\rangle = \langle w_e \rangle v. \quad (1.17)$$

The intensity shows what mean electromagnetic energy passes through a unit area in unit time. For a linear polarized wave, the intensity is proportional to the square of electric field amplitude E_0 :

$$I = \frac{1}{2} \sqrt{\frac{\epsilon\epsilon_0}{\mu\mu_0}} E_0^2. \quad (1.18)$$

For an elliptically polarized wave, the intensity is given by

$$I = \frac{1}{2} \sqrt{\frac{\epsilon\epsilon_0}{\mu\mu_0}} (E_{x0}^2 + E_{y0}^2). \quad (1.19)$$

If a linear polarized plane wave is incident upon a body, the scattered field E_s at a large distance R from the body tends to its asymptotic form of a spherical wave ($E_s \sim 1/R$). The time average intensity of field scattered into solid angle $d\omega$ is then given by

$$dI_s = I_s(\hat{o}, \hat{i}) R^2 d\omega \quad (1.20)$$

where \hat{o} and \hat{i} are the unit vectors in the scattering and incidence directions, respectively. The ratio of dI_s to the intensity of incident wave I is called the *differential scattering cross section* $\sigma_d = dI_s / I$. In radar applications and microwave propagation modeling, the *bistatic scattering cross section* $\sigma_{bi}(\sigma\hat{o}, \hat{i}) = 4\pi\sigma_d(\hat{o}, \hat{i})$ and *backscattering cross section* $\sigma_{bs} = 4\pi\sigma_d(-\hat{i}, \hat{i})$ are often used. The power scattered by the body in all directions is given by

$$P_s = \sigma_s I, \quad (1.21)$$

$$\sigma_s = \int_{4\pi} \sigma_d d\omega \quad (1.22)$$

where σ_s is the scattering cross section. The power absorbed by the body is given by

$$P_a = \sigma_a I \quad (1.23)$$

where σ_a is the absorption cross section. The sum of the scattering and absorption cross sections is called the *extinction cross section* of the body:

$$\sigma_e = \sigma_s + \sigma_a. \quad (1.24)$$

The scattering and absorption cross sections can be found from a solution to the diffraction problem or can be estimated from *integral representations*. The equivalent current density in Maxwell's equations can be written in the form (Ishimaru, 1978):

$$\vec{j}_e = \begin{cases} -j\omega\epsilon_0(\epsilon - 1)\vec{E}, & \text{inside the body} \\ 0, & \text{outside the body} \end{cases} \quad (1.25)$$

where ϵ is the complex dielectric permittivity of the body. The solution of Maxwell's equations is given by

$$\vec{E}(\vec{r}) = \vec{E}_i(\vec{r}) + \vec{E}_s(\vec{r}) \quad (1.26)$$

where $\vec{E}_i(\vec{r})$ is the incident field and $\vec{E}_s(\vec{r})$ is the scattered field. The scattered field can be found with the use of Hertz's vector \vec{A}_s :

$$\vec{E}_s(\vec{r}) = \vec{\nabla} \times \vec{\nabla} \times \vec{A}_s(\vec{r}), \quad (1.27)$$

$$A_s(r) = -\frac{1}{j\omega\epsilon_0} \int_V G_0(\vec{r}, \vec{r}') \vec{j}_e(\vec{r}') dV' = \int_V [\epsilon(\vec{r}') - 1] \vec{E}(\vec{r}') G_0(\vec{r}', \vec{r}') dV' \quad (1.28)$$

where $G_0(\vec{r}, \vec{r}') = \exp(jk_0|\vec{r} - \vec{r}'|)/(4\pi|\vec{r} - \vec{r}'|)$ is the free space Green function, k_0 is the wave vector in the free space, and $\vec{r} \neq \vec{r}'$. From (1.26)-(1.28), we obtain

$$\begin{aligned} \vec{E}(\vec{r}) &= \vec{E}_i(\vec{r}) + \vec{E}_s(\vec{r}) \\ &= \vec{E}_i(\vec{r}) + \frac{k_0^2}{4\pi} \int_V \left\{ \vec{E}(\vec{r}') \left(1 - \frac{j}{k_0|\vec{r} - \vec{r}'|} - \frac{1}{(k_0|\vec{r} - \vec{r}'|)^2} \right) - \left(\vec{E}(\vec{r}') \cdot \frac{\vec{r} - \vec{r}'}{|\vec{r} - \vec{r}'|} \right) \right. \\ &\quad \left. \times (\vec{r} - \vec{r}') \left(1 + \frac{3j}{k_0|\vec{r} - \vec{r}'|} - \frac{3}{(k_0|\vec{r} - \vec{r}'|)^2} \right) \right\} [\epsilon(\vec{r}') - 1] \frac{e^{jk_0|\vec{r} - \vec{r}'|}}{|\vec{r} - \vec{r}'|} d\vec{r}' \end{aligned} \quad (1.29)$$

Equation (1.29) represents the integral form of Maxwell's equations. It allows one, in principle, to find the electric field in any point if the spatial distribution of dielectric permittivity is given. To find the scattered field in the far-field zone ($k_0|\vec{r} - \vec{r}'| \gg 1$), at a distance R from the body, one can make the following substitutions:

$$\frac{1}{|\vec{r} - \vec{r}'|} \approx \frac{1}{R}, \quad |\vec{r} - \vec{r}'| \approx R - \vec{r}' \cdot \hat{o}. \quad (1.30)$$

The scattered field in the far-field zone is then presented as

$$\vec{E}_s(\vec{r}) = \vec{f}(\hat{o}, \hat{i}) \frac{\exp(jk_0R)}{R}, \quad (1.31)$$

$$\vec{f}(\hat{o}, \hat{i}) = \frac{k_0^2}{4\pi} \int_V [\vec{E} - \hat{o} \cdot (\hat{o} \cdot \vec{E})] [\epsilon - 1] \exp(-jk_0\vec{r}' \cdot \hat{o}) dV' \quad (1.32)$$

where $f(o, \hat{i})$ is the *scattering amplitude*. Equation (1.32) gives an explicit form for the scattering amplitude through the field inside a particle. Generally speaking, this field is unknown. However, in modeling, this field can be approximated by a known function that allows one to obtain useful approximations for the scattering amplitude.

The *optical theorem* asserts that the extinction cross section of a particle is linked with the scattering amplitude as

$$\sigma_e = \frac{4\pi}{k_0} \text{Im}[\vec{f}(\hat{i}, \hat{i}) \cdot \hat{e}_i] \quad (1.33)$$

where Im denotes the imaginary part and \hat{e}_i is the unit vector characterizing the polarization of an incident wave.

If a linear polarized wave is incident from free space upon a half-space with the flat boundary, and the wave vector of the incident wave makes an angle ϑ with the normal to the interface, the *law of refraction* is written as

$$\sin \vartheta = \sqrt{\varepsilon\mu} \sin r \quad (1.34)$$

where r is the refraction angle, ε and μ are the permittivity and permeability of the half-space. The amplitudes of reflected and refracted fields are linked with the amplitude of the incident field by *Fresnel formulas*. For a vertically polarized incident wave, these amplitudes are given by

$$E_r = -E_0 \frac{\text{tg}(\vartheta - r)}{\text{tg}(\vartheta + r)}, \quad (1.35)$$

$$E_t = E_0 \frac{2 \cos \vartheta \sin r}{\sin(\vartheta + r) \cos(\vartheta - r)} \quad (1.36)$$

where E_0 , E_r , and E_t are the amplitudes of incident, reflected, and refracted waves, respectively. For a horizontally polarized incident wave, these amplitudes are given by

$$E_r = -E_0 \frac{\sin(\vartheta - r)}{\sin(\vartheta + r)}, \quad (1.37)$$

$$E_t = E_0 \frac{2 \cos \vartheta \sin r}{\sin(\vartheta + r)}. \quad (1.38)$$

The *reflectivity* of the half-space is defined as the ratio of the reflected wave intensity to the incident wave intensity. The *transmissivity* of the half-space is defined as the ratio of the refracted wave intensity to the incident wave intensity.

If a linear polarized wave is incident upon a flat, homogeneous layer with a relative dielectric permittivity ε and thickness d , the reflectivity and transmissivity of the layer, r and t , are given by

$$r_{v,h} = \left| \frac{R_{v,h}(1 - \exp\{-j2k_{z1}d\})}{(1 - R_{v,h}^2 \exp\{-j2k_{z1}d\})} \right|^2, \quad (1.39)$$

$$t_h = \left| \frac{4k_{z0}k_{z1} \exp\{j(k_{z0} - k_{z1})d\}}{(k_{z0} + k_{z1})^2 (1 - R_h^2 \exp\{-j2k_{z1}d\})} \right|^2, \quad (1.40)$$

$$t_v = \left| \frac{4\varepsilon k_{z0}k_{z1} \exp\{j(k_{z0} - k_{z1})d\}}{(\varepsilon k_{z0} + k_{z1})^2 (1 - R_v^2 \exp\{-j2k_{z1}d\})} \right|^2 \quad (1.41)$$

where

$$k_0 = \omega/c,$$

$$k_{z0} = k_0 \cos \vartheta,$$

$$k_{z1} = k_0 \sqrt{\varepsilon - \sin^2 \vartheta},$$

$$R_h = (k_{z0} - k_{z1})/(k_{z0} + k_{z1}),$$

$$R_v = (\varepsilon k_{z0} - k_{z1})/(\varepsilon k_{z0} + k_{z1})$$

where ϑ is the incidence angle relative to the layer normal.

There is another approach to describe wave propagation in a scattering medium that is not based on Maxwell's equations but rather on energy transfer. This is called the *transfer theory*. This theory does not originate from the

wave equations (1.8), (1.9) but operates with the energy transfer in a medium containing scattering particles. This theory is built heuristically and is not rigorous in a mathematical respect. Even if the diffraction and interference effects are accounted for in the consideration of scattering by a single scatterer, the theory itself does not include diffraction effects. It is assumed in the theory that, in the summation of fields, the correlation between the fields is absent, thus, the intensities of the fields are summed but not the fields. The transfer theory operates with *ray intensity* (or *brightness*) that is the measure of energy flux. This quantity is introduced for a given direction \hat{s} as the mean energy flux density in unit spectral interval and unit solid angle. The ray intensity $J(\vec{r}, \hat{s})$ is measured in $\text{W m}^{-2} \text{Hz}^{-1} \text{sr}^{-1}$. The transfer theory takes into account the polarization effects by introducing the Stokes matrix. The vector transfer equation is of the following form: For $0 \leq \theta \leq \pi, 0 \leq \phi \leq 2\pi$,

$$\begin{aligned} \cos\theta \frac{d}{dz} \bar{I}(\theta, \phi, z) = & -\hat{\kappa}_e(\theta, \phi) \cdot \bar{I}(\theta, \phi, z) + \int_0^{2\pi} d\phi' \times \\ & \times \int_0^\pi d\theta' \sin\theta' \hat{P}(\theta, \phi; \theta', \phi') \cdot \bar{I}(\theta', \phi', z') \end{aligned} \quad (1.42)$$

where $\bar{I}(\theta, \phi, z)$ is a 4×1 column vector denoting the modified Stokes parameters in direction (θ, ϕ) , $\hat{P}(\theta, \phi; \theta', \phi')$ is the 4×4 phase matrix denoting scattering from direction (θ', ϕ') into direction (θ, ϕ) , and $\hat{\kappa}_e(\theta, \phi)$ is the extinction matrix which can be expressed in terms of the forward scattering amplitudes. For the case of thermal emission, to the right side of the equation is added the term $\hat{\kappa}_a \frac{2k_0 T(z)}{\lambda^2}$ where $\hat{\kappa}_a$ is the absorption matrix and T is the physical temperature of the medium. The vector transfer equation is usually solved numerically with the use of different approaches.

1.1.2. Blackbody Thermal Radiation

All substances radiate a continuum of electromagnetic energy owing to atomic and molecular vibrations. The nature and intensity of these vibrations depend on their temperature (thermal level of energy). The energy distribution of heated bodies' thermal emission over frequency spectrum was found with the use of the "blackbody" concept and of quantum theory assumptions.

A blackbody is an idealized body capable of absorbing all the radiation falling upon it at all frequencies, reflecting none. Furthermore, in addition to being a perfect absorber, it has to be a perfect emitter since energy absorbed

by a material would increase its temperature if no energy were emitted. The blackbody emission problem was solved by Planck. His law of radiation states that the brightness B of a blackbody radiator at a temperature T and frequency f is given by

$$B = \frac{2hf^3}{c^2} \frac{1}{e^{hf/k_B T} - 1} \quad (1.43)$$

where h is Planck's constant, k_B is Boltzmann's constant, and c is the velocity of light. The blackbody brightness is a function of only frequency and temperature, and is independent of direction and position. At low frequencies, where $hf/k_B T \ll 1$, equation (1.43) reduces to

$$B = \frac{2f^2 k_B T}{c^2} = \frac{2k_B T}{\lambda^2} \quad (1.44)$$

where $\lambda = c/f$ is the wavelength. Equation (1.44) is known as the Rayleigh-Jeans radiation formula. At a room temperature of 300 K, the fractional deviation of this formula from Planck's is less than one percent for wavelengths satisfying the condition: $\lambda > 2.57$ mm ($f \approx 117$ GHz). This covers the entire radio region and the usable part of the microwave spectrum.

In keeping with the nomenclature used to describe an ideal absorber-emitter, material objects are sometimes referred to as grey bodies. Since a blackbody radiates more energy at a given temperature than any other material, the power radiated by such a material (grey body) may be thought of as the power radiated by an equivalent blackbody having a cooler temperature than the physical temperature of the material T . Such a temperature is termed as the apparent or brightness temperature T_b of the material. The emissivity of grey body e is defined as the ratio of the body's brightness to the brightness of a blackbody of the same physical temperature. If the Rayleigh-Jeans approximation is valid, the brightness temperature of grey body relates to its temperature as

$$T_b = eT. \quad (1.45)$$

From the above definitions it follows that a blackbody has a brightness temperature identical to its physical temperature and hence an emissivity of unity. Good approximations to ideal blackbodies at microwave frequencies are the highly absorbing materials; emissivities as high as 0.99 can be achieved over a limited range of frequencies and incidence angles (relative

to normal). On the other extreme, a highly conductive metal plate is a perfect reflector with an emissivity close to zero.

Kirchhoff's law establishes the relation between the abilities of emitting and absorbing the electromagnetic energy by any physical body. In its most general formulation, this law states that the emissivity of a grey body is defined by its absorptive properties, i.e., the emissivity is equal to the absorption factor of the body.

1.1.3. Generalized Kirchhoff's Law

The thermal radiation appears as the result of random motion of charged particles inside a body. The fluctuation thermal electromagnetic field can be described as the field excited by random currents with the density $\vec{j}_e(\vec{r})$. The mean value of this density is equal to zero, and the spatial correlation function of its frequency spectrum is defined on the basis of the fluctuation-dissipation theorem. In the radio frequency band, the spatial correlation function is written as (Rytov, 1953)

$$\langle j_\alpha(\omega, \vec{r}') j_\beta(\omega, \vec{r}'') \rangle = \frac{\omega}{4\pi^2} \varepsilon''(\omega) k_B T \delta(\vec{r}' - \vec{r}'') \delta_{\alpha\beta} \quad (1.46)$$

where ε'' is the imaginary part of the body's dielectric permittivity, $\delta(\vec{r}' - \vec{r}'')$ is the delta function, and $\delta_{\alpha\beta} = 1$ ($\alpha = \beta$), $\delta_{\alpha\beta} = 0$ ($\alpha \neq \beta$). In order to calculate the fields generated by fluctuation currents, one needs to know the Green function of the considered body, i.e., the diffraction field excited inside the body by a single current source:

$$\vec{j}_G(\vec{r}') = \hat{e} \delta(\vec{r} - \vec{r}') \quad (1.47)$$

where \hat{e} is a unit vector. The field \vec{E}_d , excited by the current, is the diffraction field. To determine the fluctuation field, one can use the reciprocity theorem. Levin and Rytov (1967) have found that the fluctuation field is expressed as

$$\hat{e} \cdot \vec{E}(\vec{r}') = \int_V \vec{j}_e(\vec{r}) \vec{E}_d(\vec{r}, \hat{e}, \vec{r}') d^3\vec{r} \quad (1.48)$$

where integration is performed over the body's volume. The mean intensity of the fluctuation field component is found from equations (1.46), (1.48):

$$\left\langle \left| \hat{e} \cdot \bar{E}(\bar{r}') \right|^2 \right\rangle = \frac{\omega k_B T}{4\pi^2} \int_V \epsilon'' \left| \bar{E}_d(\bar{r}, \hat{e}, \bar{r}') \right|^2 d^3\bar{r}. \quad (1.49)$$

The expression under the integral is connected to the density of electromagnetic energy losses in the body. So, equation (1.49) can be written in the form

$$\left\langle \left| \hat{e} \cdot \bar{E}(\bar{r}') \right|^2 \right\rangle = \frac{2k_B T}{\pi} Q(\hat{e}, \bar{r}') \quad (1.50)$$

where $Q(\hat{e}, \bar{r}')$ represents the active thermal losses of the diffraction field in the body. Levin-Rytov's theory shows that to find the intensity of the fluctuation field at a given point, that is due to body's thermal emission, it is necessary to determine the thermal losses of the diffraction field exited inside the body by a unit current placed at this point and directed according to the direction of the sought fluctuation field. This approach does not apply any restrictions to the size of the body as compared to the wavelength and therefore represents a generalized form of Kirchhoff's law.

1.2. MICROWAVE RADIOMETERS AND SYSTEMS

Any microwave radiometer includes a receiving antenna and a radiometer receiver. The microwave radiometric measuring process is illustrated in Fig. 1.1.

A radiometer receiver is connected to a pencil-beam antenna pointed at a scene of interest. The objective is to measure the thermal radiation emitted by the area covered by the footprint of the antenna effective main-beam; i.e., it is desired to relate the radiometer output (represented by V_0) to the target emission that is represented by its brightness temperature T_b . To perform its function, the desired association antenna and radiometer receiver properties should be quantified.

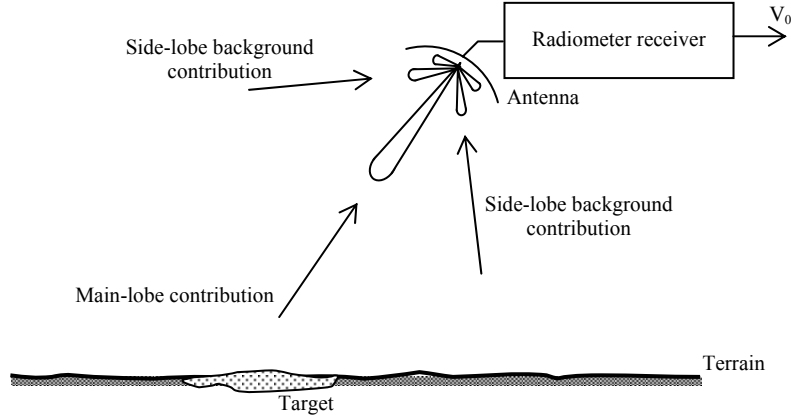


Fig. 1.1. Schematic representation of radiometric measurements.

The normalized antenna directional pattern $G(\vartheta, \varphi)$ is the angular distribution of the radiation energy flux density of the antenna in the far-field zone. For antennas of complicated configuration the antenna pattern has a multi-lobe structure. The antenna main-beam efficiency α_m and stray factor α_s are determined as follows:

$$\alpha_m = \frac{\int_{\Omega_m} G(\vartheta, \varphi) d\Omega}{\int_{4\pi} G(\vartheta, \varphi) d\Omega}, \quad (1.51)$$

$$\alpha_s = \frac{\int_{\Omega_s} G(\vartheta, \varphi) d\Omega}{\int_{4\pi} G(\vartheta, \varphi) d\Omega} = 1 - \alpha_m \quad (1.52)$$

where Ω_m is the solid angle subtended by the main lobe of the directional pattern and $\Omega_s = 4\pi - \Omega_m$. The quantity $\int_{4\pi} G(\vartheta, \varphi) d\Omega = \Omega_A$ is known as the antenna beam solid angle. The antenna directivity is defined as

$$\text{Directivity} = 4\pi / \Omega_A. \quad (1.53)$$

The other important antenna parameter is the antenna's effective area A_e . It is connected with the antenna beam solid angle as

$$A_e = \frac{\lambda^2}{\Omega_A}. \quad (1.54)$$

The linear angular resolution of an antenna is estimated by the formula

$$\theta_{3dB} \approx 1.22 \frac{\lambda}{D_A} \quad (1.55)$$

where D_A is the linear geometric size of the antenna.

The power received by a lossless microwave antenna placed inside a blackbody in a narrow bandwidth Δf is given by

$$P_A = k_B T \Delta f \frac{A_e}{\lambda^2} \int_{4\pi} G(\vartheta, \varphi) d\Omega = k_B T \Delta f \quad (1.56)$$

where T is the blackbody and antenna radiation resistance temperature. At the same time, the noise power from a resistor R at temperature T_R is given by the Nyquist formula

$$P = k_B T_R \Delta f. \quad (1.57)$$

That is identical to (1.55) if the temperature of antenna radiation resistance is T_R . Due to this identity we can express the power received by an antenna on a temperature scale. Thus the power received by the antenna is expressed in terms of antenna temperature that is defined as

$$T_A = \frac{P_A}{k_B \Delta f}. \quad (1.58)$$

The antenna temperature for an antenna with losses is expressed as

$$T_A = \frac{\eta}{4\pi} \int_{4\pi} T_b(\vartheta, \varphi) G(\vartheta, \varphi) d\Omega + (1-\eta)T_0 \quad (1.59)$$

where $T_b(\vartheta, \varphi)$ is the brightness temperature in the (ϑ, φ) direction, T_0 is the temperature of the medium surrounding the antenna, and η is the antenna radiation efficiency. Separating the power received in the main lobe and side lobes, one can obtain the other expression for the antenna temperature:

$$T_A = T_{bm} \alpha_m \eta + T_{bs} (1 - \alpha_m) \eta + (1 - \eta) T_0 \quad (1.60)$$

where T_{bm} and T_{bs} are the weighted average brightness temperatures for the antenna main and side lobes. The radiometer output voltage V_0 can be calibrated to read the antenna temperature. Equation (1.60) provides a basis for determination of the main-lobe brightness temperature from measured values of antenna temperatures:

$$T_{bm} = \frac{T_A}{\alpha_m \eta} - \frac{T_{bs} (1 - \alpha_m)}{\alpha_m} - \frac{(1 - \eta) T_0}{\alpha_m \eta}. \quad (1.61)$$

The basic function of a radiometer receiver is to measure the radiant power delivered by the antenna. For a receiver with bandwidth Δf , the power available at its input is expressed in terms of the antenna temperature as

$$P_s = k_B \Delta f \left[\frac{T_A}{L} + \left(1 - \frac{1}{L}\right) T_0 \right] \quad (1.62)$$

where L is the loss factor accounting for ohmic losses by the antenna and transmission line. The term in square brackets in (1.62) is considered as the receiver signal temperature T_s . Recovering the antenna temperature from the measured power requires precise knowledge of L and T_0 . To avoid the influence of occasional change of the loss factor and the ambient temperature, modern radiometers are designed as a single thermostatic block containing both antenna and radiometer receiver.

With no input signal present, the noise power appearing at the receiver output (prior to detection) is assigned to an effective noise temperature of the receiver T_n . The total input power is the algebraic sum of the signal and receiver noise power:

$$P_i = k_B \Delta f (T_s + T_n). \quad (1.63)$$

The simplest radiometer system is the total-power receiver that consists of a pre-detection high frequency section of bandwidth Δf and gain G ; a square-law detector with output voltage proportional to the input power; a low-pass filter-integrator section; and indicator (Fig. 1.2).

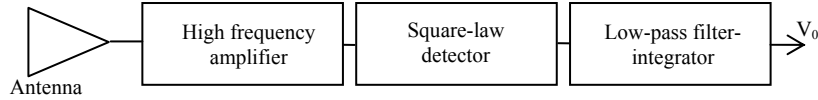


Fig. 1.2. The simplest total-power radiometer receiver.

The output voltage is related to the input power averaged over an integration time τ determined by the post-detector integrator. The detected voltage is composed of a direct current component and a fluctuating component caused by noise and gain variations. The function of the low-pass filter-integrator section is to reduce the noise variations by integrating the detector output voltage over an integration time τ . The root mean square measurement uncertainty in T_s due to noise fluctuation is estimated by

$$\Delta T_n = \frac{T_s + T_n}{\sqrt{\Delta f \tau}}. \quad (1.64)$$

Receiver gain variations contribute an additional uncertainty:

$$\Delta T_G = (T_s + T_n) \frac{\Delta G}{G} \quad (1.65)$$

where ΔG is the effective value (rms) of the detected receiver power-gain variations. These two types of variations are considered statistically independent, and therefore can be combined to define the *radiometer sensitivity*, ΔT_{\min} , as follows:

$$\Delta T_{\min} = \sqrt{(\Delta T_n)^2 + (\Delta T_G)^2} = (T_s + T_n) \sqrt{\frac{1}{\Delta f \tau} + \left(\frac{\Delta G}{G}\right)^2}. \quad (1.66)$$

Radiometer sensitivity (or radiometric resolution) is the minimum input signal variation ΔT_s required to produce a direct current change in the receiver output level equal to the root mean square value of the fluctuating component.

Dicke proposed the comparison-radiometer technique to avoid the influence of gain and receiver effective noise temperature variations. He suggested: if the receiver input is periodically switched between the antenna and a comparison, or reference, noise source at a switching rate higher than the highest significant spectral components in the gain variation spectrum, then the gain variation contribution can be reduced significantly. In other words, over a period of one switching cycle, the low-frequency (slow) gain variation component will be hardly noticeable. Since the Dicke radiometer actually measures a temperature contrast between the signal temperature and the reference temperature, its output voltage is independent of the receiver effective noise temperature. However, the desired signal temperature is now observed for only half the time (as compared to the total-power receiver). Hence, the sensitivity of this radiometer (the Dicke radiometer is called also the modulation radiometer) is

$$\Delta T_n \approx 2 \frac{T_s + T_n}{\sqrt{\Delta f \tau}}. \quad (1.67)$$

There are several other types of radiometers. Their description and parameters can be found in the literature.

To obtain a radiometric image of a scene of interest, it is necessary to scan with the antenna main-beam. If the radiometer is stationary relative to the scene, line-by-line scanning is usually used. With moving airborne platforms, scanning in only the cross-track dimension can suffice. Antenna scanning, or beam steering, can be accomplished either mechanically or electronically. Mechanical beam steering involves changing the pointing direction of the antenna axis by moving (in angle) the entire antenna or its feed. Electronic beam steering is achieved by using a planar-phased array consisting of many radiating elements. The main beam can be steered in both dimensions by electronically controlling the relative phase of each element. Radiometric systems with electronic scanning have some drawbacks: complexity, losses in the phase shifters, and high cost and weight. A simpler solution is a system that includes several antennas (push-broom systems). The antennas could be conically directed to the scene providing the same observation angles for all antennas. Also a one phase array antenna can be used to obtain several main beams by connecting antenna elements with certain constant phase shift. A separate radiometer receiver can be used for every