Traveling-Wave Slot Antennas*

J. N. HINES†, ASSOCIATE, IRE, V. H. RUMSEY†, SENIOR MEMBER, IRE, AND C. H. WALTER†, ASSOCIATE, IRE

Summary—The traveling-wave slot antenna is similar to a traveling-wave wire antenna, but it is far more versatile because the phase velocity and rate of radiation of the fields in the antenna can be controlled. Four types of traveling-wave slot antennas have been identified. These are: (a) the conventional transverse electric, TE (no tangential $E$ field parallel to the slot length); (b) transverse magnetic, TM (no tangential $H$ field parallel to the slot length); (c) a hybrid with negligible transverse $E$, and (d) a hybrid with negligible normal $H$. Only the hybrid types are capable of producing maximum radiation in the direction of the slot axis (i.e., end-fire radiation). The complex propagation constant which is characteristic of uniform traveling-wave slots has been measured for a variety of waveguide geometries and is presented in the form of graphs.

The radiation pattern of a traveling-wave slot can be controlled to give low side lobe or "cosine squared" type patterns by appropriate variation of slot width with distance $z$ along the axis. An approximate formula for the variation of attenuation $\alpha$ with $z$ required to give a specified pattern can be derived. This in turn gives the required variation of slot width with $z$.

An examination of the principle of superposition shows that the conventional technique of array design is an approximation which has proved inadequate in the design of certain slot arrays. A more elaborate technique is described which has the merit that the array pattern can be predicted exactly from measurements made by exciting individual elements of the array.

INTRODUCTION

THE APPLICATIONS of half-wave slot antennas are well-known.†-§ A traveling-wave slot antenna may be considered as a slot, many wavelengths long, which is energized at one end so that the field distribution in the slot consists of a traveling wave, in a manner analogous to traveling-wave wire antennas. It therefore performs like an array of half-wave slot antennas, but it is often much easier to design and construct than such an array. One side of a traveling-wave slot antenna is usually enclosed by a waveguide. In this form it is more versatile than the traveling-wave wire antenna because the phase velocity and rate of radiation can be controlled by the waveguide dimensions.

Various forms of traveling-wave slot antennas have been considered by different authors.†-‖ The basic features can be illustrated by considering the form suggested by Booker. In a conventional rectangular waveguide operating in the lowest order TE mode, as in Fig. 1, the magnetic field components $H_x$ and $H_z$ are zero on the central plane represented by a dashed line in Fig. 1.

It follows that this central plane can be replaced by a sheet of infinite impedance, i.e., a perfect magnetic conductor, without disturbing the field, (Fig. 2.) The equivalent of a magnetic conductor is obtained approximately when the opening in the half-guide is

* Decimal classification: R326.81. Original manuscript received by the Institute, November 10, 1952; revised manuscript received May 28, 1953. The work reported in this paper was done at the Antenna Laboratory under sponsorship of the Air Research and Development Command, Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio.
† The Antenna Laboratory, Department of Electrical Engineering, The Ohio State University Research Foundation, Columbus, Ohio.

¶ Interim Engineering Report 301-9, Antenna Laboratory, The Ohio State University Research Foundation; prepared under Contract W 33-038 ac 16520 (17380), with Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio; August 1, 1948.
fitted onto an infinite slot in a ground plane as in Fig. 3, provided the width of the slot, \( W \), is a small fraction of a wavelength. Such an arrangement, therefore, sets up a wave in an infinite slot which travels with a phase velocity very nearly the same as would be obtained in the complete waveguide.

![Fig. 3](image-url)

As the wave travels along the slot its strength is gradually dissipated by radiation. Assuming, for greater generality, that the waveguide is filled with dielectric, we can picture the radiation as due to a wave incident at angle \( \phi \) on the slot, part of which is refracted at angle \( \theta \), and the remainder reflected at angle \( \phi \) (Fig. 4).

![Fig. 4](image-url)

The refracted wave is radiated in a radiation pattern which has its main beam in the direction \( \theta \). In waveguide terminology the rate of radiation is a maximum at the cut-off frequency \( (\phi = \theta = 0) \) and diminishes to a very small value when \( \theta = 90^\circ \). (A literal interpretation of this model shows that the rate of radiation is zero when \( \theta = 90^\circ \), and for all \( \phi \) greater than the value which gives \( \theta = 90^\circ \). This result is incorrect because the actual field is not exactly the same as the unperturbed waveguide field.)

**Polarization Properties**

In a conventional slot antenna the electric field in the slot is transverse to the length because such slots are usually too narrow to support any other field configuration. If the slot is half a wavelength or more wide, this restriction disappears and it is possible to set up an arbitrary polarization in the slot.

When energy radiates from a slot (or hole of any shape) in some metal surface, the radiation field is the same as would be obtained if the slot were covered with metal and an array of magnetic currents \( K \) set up in place of the slot, where

\[
K = E \times n. \quad (1)
\]

\( E \) is the electric field of the slot and \( n \) is a unit outward vector normal to the metal surface. (This follows from Schelkunoff's equivalence principle.) When the metal surface is an infinite plane, the field is the same as the field of an array of magnetic currents \( 2K \) radiating in free space.

To illustrate this point let a set of axes be chosen as in Fig. 5. The \( x \)-axis is normal to the plane, and the \( z \)-axis is parallel to the length of the slot. Then the distribution of magnetic currents can be divided into currents flowing parallel to the \( z \)-axis, which are determined from \( E_x \) in the slot, and currents flowing parallel to the \( y \)-axis, which arise from \( E_y \).

![Fig. 5](image-url)

The electric field radiated by an individual magnetic current element lies parallel to the circles of latitude about the axis of the current element. Thus in the spherical co-ordinates \( (r, \theta, \phi) \) (Fig. 5), a magnetic current element \( K_x \) flowing parallel to the \( z \)-axis radiates an electric field which has an \( E_x \) component only. Similarly, a magnetic current element \( K_y \) radiates an electric field which has an \( E_y \) component only, where spherical co-ordinates \( (r, \xi, \eta) \) are defined in Fig. 5 also.

If a receiving dipole is oriented to pick up \( E_x \), the signal it receives must be due to only those magnetic currents which flow parallel to the \( y \)-axis. A measurement of the \( E_x \) radiation pattern over a complete hemisphere, therefore, gives all the information about \( E_x \) in the slot required to predict the radiation field. In the same way a measurement of the \( E_z \) radiation pattern can be analyzed to give the \( E_y \) distribution in the slot. The strength of \( E_y \) in the slot, relative to \( E_x \) in the slot, is given by polarization measurements in the radiation field as explained below.
When the field configuration in the slot is traveling at a constant phase velocity, the analysis of these measurements is straightforward. In this way it is possible to determine tangential E in the slot without having to set up probing apparatus in the slot (which is difficult to arrange without upsetting the field to be measured). Assume that the tangential electric field in the slot can be represented approximately by

\[ E_y = A \cos \frac{\pi y}{W} e^{-\gamma t}, \]  
\[ E_x = C \sin \frac{\pi y}{W} e^{-\gamma t}. \]

Then it can be shown that\(^{11}\)

\[ \frac{C}{A} = j \frac{\lambda}{2W} \left[ \frac{E_y/E_x}{\tan \phi} + \cos \theta \right], \]

where \(\theta\) and \(\phi\) are the co-ordinates of the point of observation and \(E_y/E_x\) is the relative magnitude and phase of the conventional far field components. The complex number \(E_y/E_x\) is obtained from the measured polarization "dumbbell." (A convenient method is described in an I.R.E. article on transmission between elliptically polarized antennas.\(^{12}\))

From such polarization measurements it has been found that there are four typical field configurations associated with traveling-wave slot antennas which are excited by a uniform waveguide whose axis is parallel to the length of the slot. These are: (a) the conventional transverse electric (no tangential \(E\) parallel to slot length);\(^{13}\) (b) transverse magnetic (no tangential \(H\) parallel to slot length); (c) a hybrid with a negligible \(E_y\) component; and (d) a hybrid with a negligible \(H_x\) component.

Any of these configurations can be excited by proper positioning of a long slot in the wall of a uniform waveguide of rectangular cross section. To obtain the transverse electric excitation a narrow slot must be placed where the current in the wall of the waveguide is perpendicular to the slot. The transverse magnetic excitation is obtained from an air-filled waveguide having a wide slot (about \(\lambda/2\)) positioned so that the current flow is parallel to the slot length. The hybrid excitation (c) is obtained under the same conditions as the TM excitation, except that the waveguide is filled with dielectric having a dielectric constant of 2.0 or greater. The hybrid excitation (d) is obtained from a waveguide partly filled with dielectric (such as paraffin) which is excited in the lowest order (hybrid) waveguide mode.

The polarization patterns show that the far field is essentially linearly polarized for \(0.4 \leq W/\lambda \leq 0.9, 1 \leq \varepsilon_r \leq 2.5\) and \(0.4 \leq \lambda/\lambda_y \leq 1.0\). In some instances slight elliptical polarization was observed. This, however, was attributed to the presence of another wave, relatively small in magnitude, traveling at a velocity close to the dominant wave. The spurious wave was presumably caused by imperfections in the antenna model used for the experiment.

**The Propagation Constant**

If the field in the slot consists of a traveling wave, the performance of the antenna can be represented by a propagation constant \(\gamma = \alpha + j\beta_n\), where \(\alpha\) and \(\beta_n\) are real. The problem is then essentially two dimensional in exactly the same sense that a conventional waveguide problem is two dimensional. If the region in which the field exists is homogeneous, the transverse propagation constant \(\kappa\) (which is related to the "free space" propagation constant \(\beta\) and to \(\gamma\) by \(\kappa^2 = \beta^2 + \gamma^2\)) is independent of frequency, and is determined uniquely by the shape of the boundary. Thus for the cross section of Fig. 3, \(\kappa\) is a function of the dimensions \(W\) and \(D\). From dimensional analysis the relation can be written in terms of two dimensionless products, as

\[ \kappa W = f\left(\frac{W}{D}\right). \]

This kind of relation is an identifying property of traveling-wave fields and can be used as a check on experimental results. The validity of this assumption has been tested by probing the field variation along the length of the slot and by measurement of the radiation pattern. It has been found that the representation by means of a single propagation constant is valid for phase velocities greater than the velocity of light.

Exploration of the field along the axis of the waveguide gives values of the attenuation constant \(\alpha\) with good accuracy except for very small values of \(\alpha\). The radiation pattern from a finite length of slot usually consists of two well-defined beams corresponding to the outgoing and reflected traveling waves, as illustrated by Fig. 6. The angle between these beams gives the phase constant \(\beta_n\), and the ratio of their amplitudes gives the attenuation constant \(\alpha\), with good accuracy provided \(\alpha\) is not too large.

Combining these techniques we have obtained experimental curves for \(\beta_n\) and \(\alpha\) which have an estimated accuracy of about 5 per cent. The results are shown in Figs. 7–14. \(\beta_n\) is related to the velocity ratio \(c/v\) by

\[ \beta_n = \frac{2\pi}{\lambda} \frac{c}{v}, \]

where \(c/v\) is the ratio of the velocity of propagation in
free space to the phase velocity along the slot. Some comparisons with theoretical results\textsuperscript{14,15} are given.

**Pattern Control**

Ideally, the pattern of a traveling-wave line source (along the $z$-axis) is obtained from source distribution function, $e^{-\pi z^2}$, multiplied by a sinusoidal factor which depends on the polarization of the source. This idealized pattern is a satisfactory approximation for practical traveling-wave slots, provided the slot is relatively narrow and an absorber is placed at the end of the slot to absorb the reflected wave. However, there are some patterns which look like the superposition of the idealized pattern and the pattern due to point sources at the ends of the slot. It has been found experimentally that this effect, when it occurs, can be greatly improved for air-filled slots by tapering the slot at the ends. The “discontinuity effect” appears to be rather unpredictable; sometimes it is insignificant in cases where on the basis of the geometrical discontinuity, one might expect it to be strongest. The indications are that it can be minimized by feeding the slot so that the mode in the feed section is as close as possible to the field configuration in the vicinity of the radiating portion.

In practice the kind of pattern control required is, for example, an approximation to a “cosecant squared” or a pattern with good directivity and low side lobes. Such patterns can be obtained to a good approximation from traveling-wave line sources if the amplitude of source strength varies with distance along the line source in the appropriate manner. For instance, an exponential variation of amplitude gives a good approximation to a “co-

secant square” pattern,\textsuperscript{16,17} and a symmetrical Gaussian ($e^{-\pi z^2}$) distribution of amplitude gives good side lobe suppression.\textsuperscript{18,19}

A traveling-wave slot antenna can be designed to give a desired amplitude distribution by using a slot whose width $W$ varies with distance $z$ along the slot. This implies that both $\alpha$ and $\beta$ are functions of $z$, but the variation of the phase constant $\beta_w$ can be made insignificant compared with the variation of $\alpha$. See Figs. 7–14.

The radiation pattern is due to an array of sources where the strength of an individual source is proportional to the current moment $M_w$ of the equivalent dipole. In order to produce a desired pattern the first step is to determine the distribution of $M_w$ which gives the best approximation to the pattern, subject to the restriction that the phase variation of $M_w$ is that of a traveling wave. The next step is to determine how the attenuation parameter $\alpha$ must vary in order to produce the desired amplitude distribution of $M_w$.

The power radiated from a source of given current moment $M_w$ is proportional to $M_w^2$ if the source is radiating by itself. In the actual problem a given source radiates in the presence of the rest. Consequently, the power radiated from a given source is proportional to $M_w^3$ provided that the variation of source strength with position is gradual, and that contributions to the field at the source from distant sources is insignificant. These restrictions appear to be approximated in practice except for the case when the phase velocity equals $c$ where the latter condition would not be justifiable, \textit{a priori}. Under conditions where these two provisions are satisfied it is possible to write

\[ \frac{dP}{dz} \sim A^2, \]

where $P$ is the power flowing down the guide in the $z$-direction, and $A$ represents the amplitude distribution of current moment. Therefore,

\[ 2\alpha = -\frac{1}{P} \frac{dP}{dz} = -\frac{A^2}{\int A^2 dz}. \] (6)


\textsuperscript{16} Interim Engineering Reports 301-14, Section 4; February 1, 1949, and 301-19, Section 8; October 1, 1949, Antenna Laboratory, The Ohio State University Research Foundation; prepared under Contract W 33-038 AC 16520 (17380) with Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio.


\textsuperscript{18} Interim Engineering Report 301-15, Section 4, Antenna Laboratory, The Ohio State University Research Foundation; prepared under Contract W 33-038 AC 16520 (17380) with Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio; May 1, 1949.

\textsuperscript{19} Interim Engineering Report 301-16, Section 4, Antenna Laboratory, The Ohio State University Research Foundation; prepared under Contract W 33-038 AC 16520 (17380) with Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio; August 1, 1949.
Given the amplitude distribution $A$, (6) determines $\alpha$ as a function of $z$, but it does not give $\alpha(z)$ uniquely because the limits on the integral are unspecified. If the slot extends from $z = 0$ to $z = L$, the input power, $P(0)$, is related to the power left over, $P(L)$, by

$$P(0) - P(L) \sim \int_{0}^{L} A^2dz.$$  \hspace{1cm} (7)

Then (6) can be expressed in the form

$$2\alpha(z) = \frac{A^2}{\int_{0}^{L} A^2dz + \frac{P(L)}{P(0) - P(L)} \int_{0}^{L} A^2dz}.$$  \hspace{1cm} (8)

Equation (8) shows that $\alpha(z)$ is determined not only by the amplitude distribution $A$, but also by the fraction of incident power which is radiated. If practically all of the power is radiated, $\alpha$ rises to a very high value.
near the end, $z = L$. In practice one would set the radiated power one or two decibels down from the incident power and then apply (8) to determine $\alpha(z)$.

Pattern control for TM or hybrid slots is usually simpler than for TE slots because the variation of $W$ required for a given variation of $\alpha$ is much smaller for TM or hybrid operation than for TE operation. Despite this, excellent control can be obtained even with TE excitation. A Gaussian distribution was approximated without excessive phase change by placing an iris of variable width $W$ over a TE-excited guide Fig. 15.

Good side-lobe suppression was obtained over $2:1$ frequency range, the side lobes being insignificant over a substantial part of the range. Typical radiation patterns are shown in Fig. 16.

The attenuation $\alpha$ can be alternatively controlled by means of a horn fitted onto the slot so that the horn acts as a transformer between the slot aperture and free space. In this way successful side-lobe suppression can be obtained by varying the length of horn to produce the desired amplitude distribution.$^{20}$

$^{20}$ Interim Engineering Report 301-19, Section 8, Antenna Laboratory, The Ohio State University Research Foundation; prepared under Contract W 53-038 ac 16520 (17380) with Wright Air Development Center, Wright-Patterson Air Force Base, Dayton, Ohio; October 1, 1949.
Arrays of Traveling-Wave Slots

The problem of designing a flush-mounted directional antenna to fit in a given area is a typical application for a traveling-wave slot array. The conventional array problem is to design a suitable element (or primary radiator) of the array and then to design a feed system for energizing an appropriate array of these elements to give the desired pattern. The pattern of the array is obtained by superimposing the individual radiation fields of the elements of the array. In the familiar case of an array of half-wave dipoles it happens that the individual radiation field of a dipole (called the primary pattern for simplicity) is practically the same whether the dipole radiates in free space or in the presence of the other dipoles, provided they are open-circuited. The preoccupation with arrays of half-wave dipoles has made this fact so familiar that one is apt to assume (erroneously) that the primary pattern is, in general, the pattern of an element of the array radiating by itself.

In order to apply the principle of superposition correctly the first step is to choose the location of the “input terminals” of each element of the array. For this purpose “input terminals” can be defined as some accessible point in the feed system where the field consists of a single mode. Then the primary pattern $F_i$ of an individual radiator can be defined as the field due to unit input current to its terminals in the presence of all the other radiators when their input terminals are open-circuited. Then the pattern of the array, when the input currents are $I_1, I_2, \ldots$, is given by $\sum_i I_i F_i$.

Alternatively, the primary pattern $G_i$ may be defined as the field due to unit input voltage to the $i^{th}$ radiator when the others are short-circuited, in which case the array pattern due to input voltages $V_1, V_2, \ldots$ is given by $\sum_i V_i G_i$. Note that the “primary pattern” $F_i$ or $G_i$ may be interpreted as a vector, having components $E_i$ and $H_i$, where $E_i$ and $H_i$ are themselves represented by complex numbers representing the radiated signal in phase and amplitude relative to the input signal $I_i$, in the case of $F_i$, or $V_i$ in the case of $G_i$.

Having determined the primary patterns $F_i$ (or $G_i$), the problem is to find the distribution of $I_i$ (or $V_i$) which gives the best approximation to the desired pattern. This problem is greatly simplified if the mutual impedance between radiators (with respect to the chosen set of input terminals) is insignificant. Then the problem of designing the feed system reduces to the problem of feeding a known set of loads with specified currents or voltages.

This technique of array design is much more involved than the conventional method but it is absolutely accurate, and it has been found from experience on traveling-wave slot arrays designed for low side lobes, that the conventional technique is inadequate.

The mutual impedance between traveling-wave slot antennas may be analyzed by picturing the traveling-wave slot as a directional coupler. When two such slots are placed along parallel lines, the field induced in one due to energization of the other will consist primarily of a wave traveling in the same direction as the wave in the slot which is energized directly. This wave will be reflected at the end of the parasitic slot and will travel back to its input terminals. However, it is attenuated by ra-
diation as it travels along the slot so that the amount available at the input terminals of the parasitic slot is very small. If the slots are long enough to be good directional antennas, the coupling between adjacent input terminals is very small. In practice it has proved to be negligible.

The magnitude of the mutual impedance is quite different for the TE, TM, or hybrid slots. For example the equivalent magnetic dipoles associated with parallel slots are broadside-to-broadside for TE operation, whereas they are end-to-end for the hybrid operation (c) (no transverse E). Consequently the mutual impedance effects associated with hybrid operation (c) are much less than with TE operation.

ACKNOWLEDGMENT

This work was performed under Contract AF 33(038)-9236 between Wright-Patterson Air Force Base and The Ohio State University Research Foundation. It is a pleasure to acknowledge the help given by R. W. Masters, O. Click, R. Krausz, B. J. Stephenson and other members of the Ohio State University Antenna Laboratory Staff.

Prediction of Traveling Wave Magnetron Frequency Characteristics: Frequency Pushing and Voltage Tuning*

H. W. WELCH, JR.,†, SENIOR MEMBER, IRE

Summary—An approximate method has been developed for determining the shape and density of the spokes of electronic space charge when large RF potentials exist in the magnetron. With the estimation of space-charge configuration which this method makes possible, induced current theory and knowledge of the magnetron electrode geometry and external circuit can be applied to the calculation of frequency characteristics. These characteristics relate frequency to operating anode potential and anode current and are defined as frequency pushing or voltage tuning characteristics. Relatively simple equations for these characteristics are presented. Calculated characteristics for typical values of the variables of magnetron design are presented. The correspondence of the results of the theory with experimental data is discussed very briefly.

INTRODUCTION

THIS PAPER PRESENTS an analysis which leads to the understanding of two cw magnetron characteristics which are important in the use of magnetrons for radio communication; namely, frequency pushing and voltage tuning. These terms require definition and part of the purpose of the following discussion will be to provide a basis for explicit definition. Definitions acceptable to those who are well acquainted with cw magnetron behavior are the following:

Frequency pushing is defined as the variation of the frequency which is generated by an oscillating magnetron and which is associated with the change in dc anode current as the anode voltage is raised with the resonator temperature held constant. A frequency-pushing characteristic would therefore be a plot of generated frequency versus dc anode current for the conditions mentioned. Actually, to be complete, a three-dimensional plot including anode voltage, current, and frequency, should be used.

A typical frequency-pushing characteristic is shown in Fig. 1a. Voltage tuning is defined as the variation in frequency which is generated by an oscillating magnetron and which is associated with the change in dc anode voltage when dc anode current, load impedance, magnetic field, and resonator temperature are held constant. In order for dc anode current to be held constant, the cathode emission must be limited in some way, such as by operation at a reduced temperature. Otherwise, an increase in anode voltage will increase the anode current. A voltage-tuning characteristic is therefore a plot of generated frequency versus anode voltage for the conditions mentioned.

A typical voltage-tuning characteristic is shown in Fig. 1b. Examination of a number of typical sets of data on cw magnetrons shows that, in normal operation (when frequency pushing is observed), the frequency is relatively insensitive to the voltage change which is required to increase anode current compared to the large frequency shifts obtained under the conditions required for voltage-tuning operation. For example, in the figures shown, a change of 100 volts produces a frequency pushing of 10 megacycles, while a change of 12.2 volts produces a voltage tuning of the same amount. Typical voltage tuning characteristics show a change of from 0.1 to 2 megacycles per volt.

* Decimal classification: R355.912.1×R139. Original manuscript received by the Institute, February 2, 1953; revised manuscript received, June 8, 1953. (This paper is based on work done for the Signal Corps, U. S. Army, Contract No. DA-36-039 sc-5423. It is a condensed part of a thesis submitted by the author in partial fulfillment of the requirements for the Ph.D. degree at the University of Michigan. "Dynamic frequency characteristics of the magnetron space charge; frequency pushing and voltage tuning," also issued as Tech. Report No. 12, Electron Tube Lab., Dept. Elec. Eng., University of Michigan; November, 1951.

† University of Michigan, Ann Arbor, Mich.
专注于微波、射频、天线设计人才的培养
易迪拓培训
网址：http://www.edatop.com

射频和天线设计培训课程推荐

易迪拓培训(www.edatop.com)由数名来自于研发第一线的资深工程师发起成立，致力并专注于微波、射频、天线设计研发人才的培养；我们于2006年整合合并微波EDA网(www.mweda.com)，现已发展成为国内最大的微波射频和天线设计人才培养基地，成功推出多套微波射频以及天线设计经典培训课程和ADS、HFSS等专业软件使用培训课程，广受客户好评；并先后与人民邮电出版社、电子工业出版社合作出版了多本专业图书，帮助数万名工程师提升了专业技术能力。客户遍布中兴通讯、研通高频、埃威航电、国人通信等多家国内知名公司，以及台湾工业技术研究院、永业科技、全一电子等多家台湾地区企业。

易迪拓培训课程列表：http://www.edatop.com/peixun/rfe/129.html

射频工程师养成培训课程套装

该套装精选了射频专业基础培训课程、射频仿真设计培训课程和射频电路测量培训课程三个类别共30门视频培训课程和3本图书教材；旨在引领学员全面学习一个射频工程师需要熟悉、理解和掌握的专业知识和研发设计能力。通过套装的学习，能够让学员完全达到和胜任一个合格的射频工程师的要求…

课程网址：http://www.edatop.com/peixun/rfe/110.html

ADS学习培训课程套装

该套课程是迄今国内最全面、最权威的ADS培训教程，共包含10门ADS学习培训课程。课程是由具有多年ADS使用经验的微波射频与通信系统设计领域资深专家讲解，并多结合设计实例，由浅入深、详细而又全面地讲解了ADS在微波射频电路设计、通信系统设计和电磁仿真设计方面的内容。能让您在最短的时间内学会使用ADS，迅速提升个人技术能力，把ADS真正应用到实际研发工作中去，成为ADS设计专家…


HFSS学习培训课程套装

该套课程套装包含了本站全部HFSS培训课程，是迄今国内最全面、最专业的HFSS培训教程套装，可以帮助您从零开始，全面深入学习HFSS的各项功能和在多个方面的工程应用。购买套装，更可超值赠送3个月免费学习答疑，随时解答您学习过程中遇到的棘手问题，让您的HFSS学习更加轻松顺畅…

课程网址：http://www.edatop.com/peixun/hfss/11.html
专注于微波、射频、天线设计人才的培养
网址：http://www.edatop.com

**CST 学习培训课程套装**

该培训套装由易迪拓培训联合微波 EDA 网共同推出，是最全面、系统、专业的 CST 微波工作室培训课程套装，所有课程都由经验丰富的专家授课，视频教学，可以帮助您从零开始，全面系统地学习 CST 微波工作的各项功能及其在微波射频、天线设计等领域的设计应用。且购买该套装，还可超值赠送 3 个月免费学习答疑⋯


**HFSS 天线设计培训课程套装**

套装包含 6 门视频课程和 1 本图书，课程从基础讲起，内容由浅入深，理论介绍和实际操作讲解相结合，全面系统的讲解了 HFSS 天线设计的全过程。是国内最全面、最专业的 HFSS 天线设计课程，可以帮助您快速学习掌握如何使用 HFSS 设计天线，让天线设计不再难⋯

课程网址：http://www.edatop.com/peixun/hfss/122.html

**13.56MHz NFC/RFID 线圈天线设计培训课程套装**

套装包含 4 门视频培训课程，培训将 13.56MHz 线圈天线设计原理和仿真设计实践相结合，全面系统地讲解了 13.56MHz 线圈天线的工作原理、设计方法、设计考量以及使用 HFSS 和 CST 仿真分析线圈天线的具体操作，同时还介绍了 13.56MHz 线圈天线匹配电路的设计和调试。通过该套课程的学习，可以帮助您快速学习掌握 13.56MHz 线圈天线及其匹配电路的原理、设计和调试⋯


**我们的课程优势：**

※ 成立于 2004 年，10 多年丰富的行业经验，
※ 一直致力并专注于微波射频和天线设计工程师的培养，更了解该行业对人才的要求
※ 经验丰富的一线资深工程师讲授，结合实际工程案例，直观、实用、易学

**联系我们：**

※ 易迪拓培训官网：http://www.edatop.com
※ 微波 EDA 网：http://www.mweda.com
※ 官方淘宝店：http://shop36920890.taobao.com

专注于微波、射频、天线设计人才的培养
官方网址：http://www.edatop.com
淘宝网店：http://shop36920890.taobao.com