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Numerical solution to the current integral equations of a boundary-penetrating antenna*

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Abstract: Based on the current integral equations of a Boundary-Penetrating Antenna (BPA) embedded in a three-layer dissipative medium, the closed-form low frequency approximation of the Sommerfeld-type integrals involved in the current integral equations of the BPA had been deduced. The physical interpretation of the closed-form formulas was discussed. The input impedance of the BPA is calculated with the three-term current distribution assumption (entire-domain basic function). Furthermore, a simple quasi-static approximation formula was employed to obtain the electric fields on the ground due to the BPA. Finally, a numerical example was given.

Key words: boundary-penetrating antenna; sommerfeld integral; layered Medium; input impedance; drill-rod telemetry system

1 Iintroduction

When the analysis of an electromagnetic problem is concerned with the earth, the earth is often assumed simply as a homogeneous dissipative medium[1,2]. But in the analysis of a drill-rod telemetry system, because the drill-rod is very long and buried in the earth, a more precise model of the earth is needed. Therefore, the earth should be thought as a stratified medium, at least, two-layer lossy medium, one soil and one rock. In[3], a modified perturbation method has been presented to analyze the drill-rod telemetry system, in which the exciting source is a loop current. If the drill-rod itself is used as a dipole antenna excited by a δ voltage source, the analysis of the antenna shown in fig. 1 is necessary. This is a very difficult problem because the antenna, we called it 'boundary-penetrating antenna (BPA)', is so long that it stretches through the interface between the two medium. So far, there are few references

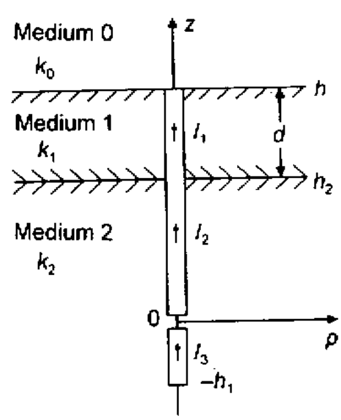

Figure1. Geometric configuration of a boundary-penetrating antenna.

report ing the study on this kind of antenna.

In [4], the current integral equations of a BPA buried in a three-layer dissipative medium have been deduced by using the Hertz vector and the principle of superposition. The current distribution and the input impedance of the antenna

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can be calculated with Method of Moments (MoM). The numerical calculation of the Sommerfeld-type integrals in the current integral equations is rather complicated and time-consuming. In this paper, the Sommerfeld-type integrals involved in the current integral equations are simplified when the frequency used is extremely low. How to get the numerical solution of the complex current integral equations is discussed. A quasi-static approximation formula is utilized to obtain the electromagnetic fields due to the antenna. At last, the input impedance of the BPA and the electric fields on the ground are calculated.

2 Current Integral Equations of BPA

In [4], the current integral equations of a boundary-penetrating antenna buried in a three-layer dissipative medium have been worked out. For convenience and the entirety of the paper, we repeat them in following

$$\frac{\mu_0}{4\pi}\left[\int_{-h_1}^{0} I_3(z')+\int_0^{h_2} I_2(z')\right]G_{21}(z,z')\mathrm{d}z' + \frac{\mu_0}{4\pi}\int_{h_2}^{h} I_1(z')G_{11}(z,z')\mathrm{d}z' = C_1\cos k_1 z + C_2\sin k_1 z \qquad h_2\le z\le h \tag{1a}$$

$$\frac{\mu_0}{4\pi}\left[\int_{-h_1}^{0} I_3(z')+\int_0^{h_2} I_2(z')\right]G_{22}(z,z')\mathrm{d}z' + \frac{\mu_0}{4\pi}\int_{h_2}^{h} I_1(z')G_{12}(z,z')\mathrm{d}z' = C_3\cos k_2 z + C_4\sin k_2 z \qquad -h_1\le z\le h_2 \tag{1b}$$

where

$$G_{11}(z,z')=\int_0^\infty \frac{\lambda}{u_1}\left[\mathrm{e}^{-u_1|z-z'|}+f_2(\lambda)e^{u_1 z}+f_2(\lambda)\mathrm{e}^{-u_1 z}\right]\mathrm{J}_0(\lambda\rho)\mathrm{d}\lambda \tag{2a}$$

$$G_{12}(z,z')=\frac{k_2^2}{k_1^2}\int_0^\infty f_4(\lambda)\mathrm{e}^{u_2 z}\mathrm{J}_0(\lambda\rho)\mathrm{d}\lambda \tag{2b}$$

$$G_{21}(z,z')=\frac{k_1^2}{k_2^2}\int_0^\infty\left[g_2(\lambda)\mathrm{e}^{u_1 z}+g_3(\lambda)\mathrm{e}^{-u_1 z}\right]\mathrm{J}_0(\lambda\rho)\mathrm{d}\lambda \tag{2c}$$

$$G_{22}(z,z')=\int_0^\infty \frac{\lambda}{u_2}\left[\mathrm{e}^{-u_2|z-z'|}+g_4(\lambda)\mathrm{e}^{u_2 z}\right]\mathrm{J}_0(\lambda\rho)\mathrm{d}\lambda \tag{2d}$$

and

$$f_2(\lambda)=\frac{-\alpha_-\beta_-\mathrm{e}^{-u_1 z'}+\alpha_-\beta_+\mathrm{e}^{u_1 z'-2u_1 h_2}}{\Delta} \tag{3a}$$

$$f_3(\lambda)=\frac{-\alpha_+\beta_-\mathrm{e}^{-u_1 z'+2u_1 h}-\alpha_-\beta_+\mathrm{e}^{u_1 z'}}{\Delta} \tag{3b}$$

$$f_4(\lambda)=\frac{2\lambda}{u_2}\frac{\alpha_-\mathrm{e}^{u_1 z'}+\alpha_+\mathrm{e}^{-u_1 z'+2u_1 h}}{\Delta}\mathrm{e}^{-(u_1+u_2)h_2} \tag{3c}$$

$$g_2(\lambda)=\frac{2\lambda}{u_1}\frac{\alpha_-(\beta u_1/u_2)\mathrm{e}^{u_2 z'}}{\Delta}\mathrm{e}^{-(u_1+u_2)h_2} \tag{3d}$$

$$g_3(\lambda)=\frac{2\lambda}{u_1}\frac{\alpha_+(\beta u_1/u_2)\mathrm{e}^{u_2 z'}}{\Delta}\mathrm{e}^{-(u_1+u_2)h_2+2u_1 h} \tag{3e}$$

$$g_4(\lambda)=\frac{\left[\alpha_-\beta_++\alpha_+\beta_-\mathrm{e}^{2u_1(h-h_2)}\right]\mathrm{e}^{u_2 z'-2u_2 h_2}}{\Delta} \tag{3f}$$

$$\Delta=\alpha_+\beta_+\mathrm{e}^{2u_1(h-h_2)}+\alpha_-\beta_- \tag{3g}$$

where $\alpha_\pm=1\pm\alpha u_0/u_1$, $\beta_\pm=1\pm\beta u_1/u_2$, $\alpha=k_1^2/k_0^2$, $\beta=k_2^2/k_1^2$, $u_i=(\lambda^2-k_i^2)^{1/2}$, $i=1,2$, $k=(\omega^2\mu_0\varepsilon-j\omega\mu_0\sigma)^{1/2}$ · $\mathrm{J}_0(\lambda\rho)$ is the zeroth order Bessel function of the 1st kind. C_1, C_2, C_3, C_4 are undetermined constants. Eqs. (1) are the current integral equations of a BPA buried in a three-layer medium.

3 Simplification of the Formulas in Extremely Low Frequency

The current integral equations of the antenna can be resolved with Method of Moments (MoM). The numerical calculation of the Sommerfeld-type integrals in eqs. (2) is rather complicated and time-consuming. In actual case, the medium 0 is air, medium 1 soil and medium 2 rock. So when the frequency is lower than 1MHz, the displacement current can be negligible. In general, the conductivity of soil and rock are $\sigma_1=10^{-2}$S/m, $\sigma_2=10^{-4}$S/m, respectively. Then we have (the detail derivation is presented in the Appendix)

$$|\alpha u_0/u_1|\gg 1,\ |\beta u_1/u_2|\ll 1 \tag{4}$$

and let $d=h-h_2$, eqs. (3) can be reduced as

$$f_2(\lambda)\approx\frac{\mathrm{e}^{-u_1 z'}-\mathrm{e}^{u_1 z'-2u_1 h_2}}{e^{2u_1 d}-1}$$

$$f_3(\lambda)\approx\frac{e^{u_1 z'}-e^{-u_1 z'+2u_1 h}}{e^{2u_1 d}-1}$$

$$f_4(\lambda)\approx\frac{2\lambda}{u_2}\frac{e^{-u_1 z'+2u_1 h}-e^{u_1 z'}}{e^{2u_1 d}-1}e^{-(u_1+u_2)h_2}$$

$$g_2(\lambda)\approx-\frac{2\lambda}{u_1}\frac{(\beta u_1/u_2)e^{u_2 z'}}{e^{2u_1 d}-1}e^{-(u_1+u_2)h_2}$$

$$g_3(\lambda) \approx \frac{2\lambda}{u_1} \frac{(\beta u_1/u_2)e^{u_2 z'}}{e^{2u_1 d}-1} e^{-(u_1+u_2)h_2+2u_1 h}$$

$$g_4(\lambda) \approx e^{u_2 z'-2u_2 h_2}$$

Noticing the Sommerfeld integral

$$\frac{e^{-jk_i(z^2+\rho^2)^{1/2}}}{(z^2+\rho^2)^{1/2}} = \int_0^\infty \frac{\lambda}{u_i} e^{-u_i|z|} \mathrm{J}_0(\lambda\rho)\mathrm{d}\lambda \tag{5}$$

we have

$$G_{22}(z,z') = \frac{e^{-jk_2 R}}{R} + \frac{e^{-jk_2 R_{21}}}{R_{21}} \tag{6}$$

where $R=[a^2+(z-z')^2]^{1/2}$,

$R_{21}=[a^2+(z+z'-2h_2)^2]^{1/2}$

Noting that $\frac{1}{1-x}=\sum_{n=0}^{\infty} x^n$ and $|e^{-2u_1 d}|<1$, $f_2(\lambda)$ can be expanded as

$$f_2(\lambda) = \sum_{n=0}^{\infty} e^{-u_1[z'+2(n+1)d]} - \sum_{n=0}^{\infty} e^{-u_1(2h-z'+2nd)}$$

similarly

$$f_3(\lambda) = \sum_{n=0}^{\infty} e^{-u_1[-z'+2(n+1)d]} - \sum_{n=0}^{\infty} e^{-u_1(z'-2h_2+2nd)}$$

then

$$G_{11}(z,z') = \frac{e^{-jk_1 R}}{R} + \sum_{n=0}^{\infty}\Big[\frac{e^{-jk_1 R_{11}}}{R_{11}} - \frac{e^{-jk_1 R_{12}}}{R_{12}} + \frac{e^{-jk_1 R_{13}}}{R_{13}} - \frac{e^{-jk_1 R_{14}}}{R_{14}}\Big] \tag{7}$$

where

$$R_{11} = [a^2 + (z - z' - 2(n+1)d)^2]^{1/2},$$
$$R_{12} = [a^2 + (z + z' - 2h - 2nd)^2]^{1/2},$$
$$R_{13} = [a^2 + (z - z' + 2(n+1)d)^2]^{1/2},$$
$$R_{14} = [a^2 + (z + z' - 2h_2 + 2nd)^2]^{1/2}$$

and is the radius of the antenna.

The physical interpretation of the above results is clear. Under the approximation of eq. (4), medium 1 can be considered a perfect conductor respect to medium 2, so the closed-formula of eq. (6) can be obtained very directly. The second term on the right side of eq. (6) indicates the image of a dipole in medium 2 reflected on the boundary plane between media 1 and 2. As we known, for a vertical dipole in the medium 1, the reflection coefficients at the upper boundary (media 0—1) and lower one (media 1—2) are both −1 [1], therefore, the infinite summation term in eq. (7) represents the infinite series of the images of the dipole reflected on both boundary planes.

$G_{12}(z,z')$, $G_{21}(z,z')$ can not be reduced to closed-form formulae. Fortunately, the calculation of them can be avoided through the discussion of the physical meanings of eqs. (1). When two dipole antennas are located in close proximity to each other, the current distribution on one is affected by the field radiated from the other one [2]. In fact, the boundary-penetrating antenna may be considered a pair of antennas: the upper part of the antenna in medium 1 and the lower part of the antenna in medium 2. The first term of the left side of eq. (1a) and the second term of the left side of eq. (1b) are the mutual terms. As we discuss above, in the condition of extremely low frequency, to a reasonable approximation, medium 1 can be considered a perfect conductor respect to medium 2. That means that the effect on the upper part of the lower part is trivial. Due to the attenuation of the lossy medium, the current on the upper part of the antenna is much small. Moreover, from the view of math, there is a factor of $\beta=k_2^2/k_1^2$, which is great less than 1, in $G_{12}(z,z')$ of eq. (2b), so the second term of the left side of eq. (1b) contributes much less to the eq. (1b). Hence both the mutual terms in eqs. (1) are negligible. But we should bear in mind, wiping off the mutual terms does not mean that the upper part and the lower part of the antenna do not affect each other. In fact, the effect of them are implied in that the currents on the upper part and the lower part must be continual at the interface between media 1 and 2.

4 Numerical Solutions and Discussion

Suppose a δ-source, V_0, is applied on the antenna, we have the current integral equation of the antenna [5]

$$\Big[\int_{-h_1}^{0} I_3(z') + \int_0^{h_2} I_2(z')\Big]G_{22}(z,z')\mathrm{d}z' = C'_1\cos k_2 z - j\frac{2\pi k_2}{\omega\mu_0}V_0\sin k_2|z| \tag{8a}$$

$$\int_{h_2}^{h_3} I_1(z')G_{11}(z,z')\mathrm{d}z' = C'_2\cos k_1 z + C'_3\sin k_1 z \tag{8b}$$

where C'_1, C'_2, C'_3 are undetermined constants. Eqs. (8) can be solved numerically to get the current distribution on the dipole antenna by using Method of Moments with the three-term current distribution assumption (entire-domain primary function)

$$I_1(z') = A_1 \sin k_1(h_3 - z') + B_1[\cos k_1(z' - h_2) - \cos k_1(h_3 - h_2)] + C_1\left[\cos\frac{k_1}{2}(z' - h_2) - \cos\frac{k_1}{2}(h_3 - h_2)\right] \quad (9a)$$

$$I_2(z') = A_2 \sin k_2(h_3 - z') + B_2(\cos k_2 z' - \cos k_2 h_3) + C_2\left(\cos\frac{k_2}{2}z' - \cos\frac{k_2}{2}h_3\right) \quad (9b)$$

$$I_3(z') = A_3 \sin k_2(-h_1 - z') + B_3(\cos k_2 z' - \cos k_2 h_1) + C_3\left(\cos\frac{k_2}{2}z' - \cos\frac{k_2}{2}h_1\right) \quad (9c)$$

Considering the current continue conditions

$$I_3(0) = I_2(0),\ I_2(h_2) = I_1(h_2) \quad (10)$$

we may let

$$A_1 = \frac{\sin k_2(h_3 - h_2)}{\sin k_1(h_3 - h_2)} A_2,$$

$$B_1 = \frac{\cos k_2 h_2 - \cos k_2 h_3}{1 - \cos k_1(h_3 - h_2)} B_2,$$

$$C_1 = \frac{\cos\frac{k_2}{2}h_2 - \cos\frac{k_2}{2}h_3}{1 - \cos\frac{k_1}{2}(h_3 - h_2)} C_2,$$

$$A_3 = -\frac{\sin k_2 h_3}{\sin k_2 h_1} A,\ B_3 = \frac{1 - \cos k_2 h_3}{1 - \cos k_2 h_1} B_2,$$

$$C_3 = \frac{1 - \cos\frac{k_2}{2}h_3}{1 - \cos\frac{k_2}{2}h_1} C_2$$

The input impedance of the antenna can be obtained from the following equation

$$Z_{in} = V_0 / I_2(0) = V_0 / I_3(0) \quad (11)$$

In general, the electromagnetic fields can be obtained by the integration of the current over the antenna when the current distribution on the antenna is available. But it is rather complex due to the existence of the layered medium. Considering the extreme low frequency, we adopt another more simple approximate approach to calculate the fields. We are just interested in the electromagnetic fields that are on the surface of the earth and near the antenna ($\rho \leq 100$m), so comparing with the 1000 meter-length of the BPA, the drill-rod may be considered as a semi-infinite stick. Then there is a quasi-static approximation formula of the electromagnetic field [6]

$$E_\rho(\rho, z) \approx \frac{\omega^2 \mu_0 q(z)}{4\pi k_1^2 \rho} \quad (12)$$

where $q(z) = \frac{j}{\omega}\frac{dI(z)}{dz}$. Also the dominant part of the current on the antenna is I_2, hence from eqs. (9) and (12) we have

$$E_p(p, z) \approx -j\frac{\omega\mu_0 k_2}{4\pi k_1^2 p}\left[A_2 \cos k_2(h - z) + B_2 \sin k_2 z + \frac{C_2}{2}\sin\frac{k_2 z}{2}\right] \quad (13)$$

For the typical parameters: $h = 1000$m, $d = 100$m, $\sigma = 10^{-2}$s/m, $\sigma_2 = 10^{-4}$s/m, $a = 0.1$m, where σ_1, σ_2, a are the conductivity of soil, rock and the radius of the antenna, respectively. We have the results shown in Figs. 2~5.

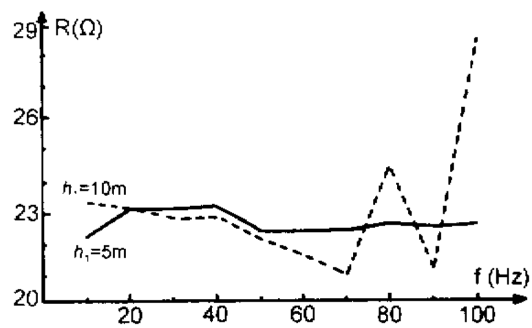
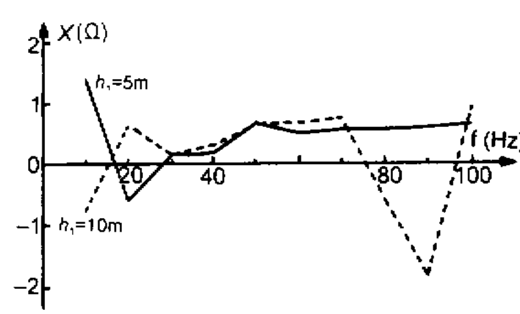


Figure2. The input resistance for the BPA versus frequency



igure3. The input reactance for the BPA versus frequency

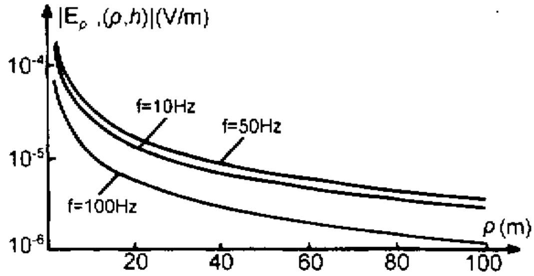


Figure4. The electric field on the surface of the ground(m, V)

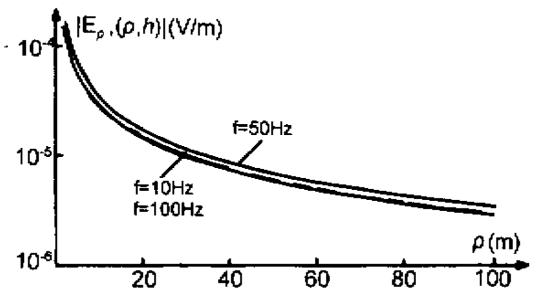


Figure5. The electric field on the surface of the ground(m, V)

From Figs. 2 and 3 we can see that the input reactance of the BPA is small, and the input resistance varies slow with the frequencies, then the match between the antenna and the exciting source is easy. Comparing with the case of a loop antenna in a borehole[3], it has shown clearly in Figs. 4 and 5 that the efficiency of a dipole antenna is much better. Consequently, from the view of application, it is obvious that the exciting sort of a dipole antenna is more practical than that of a loop antenna used in an electric drill-rod telemetry system.

The model presented in this paper is much close to the reality of the electric drill-rod telemetry system, more difficult to analyze, of course. The Sommerfeld-type integrals involved in the current integral equations of the BPA are reduced to closed-form formulas when the frequency is extremely low and the numerical computing time is saved much. The quasi-static formula leads to get the electromagnetic fields simply.

Appendix

λ is real when the integral route in Sommerfeld integrals (2) is along the real axis. In the case the displacement current is negligible, we have $k_i^2 = -j\omega\mu_0\sigma_i$, then $k_i^2/k_j^2 = \sigma_i/\sigma_j$ is real. Let

$$\frac{\sqrt{\lambda^2 + j\omega\mu_0\sigma_i}}{\sqrt{\lambda^2 + j\omega\mu_0\sigma_j}} = u + jv$$

with a not complex algebra procedure we can obtain

$$u^2 + v^2 = \frac{[\lambda^8 + \omega^2\mu_0^2(\sigma_i^2 + \sigma_j^2)\lambda^4 + \omega^4\mu_0^4\sigma_i^2\sigma_j^2]^{1/2}}{\lambda^4 + \omega^2\mu_0^2\sigma_j^2}$$

$u^2 + v^2$ is the function of λ, let $h(\lambda) = u^2 + v^2$, we have its derivative

$$h'(\lambda) = \frac{2h(\lambda)\lambda^3\omega^2\mu_0^2}{\lambda^8 + \omega^2\mu_0^2(\sigma_i^2 + \sigma_j^2)\lambda^4 + \omega^4\mu_0^4\sigma_i^2\sigma_j^2}(\sigma_j^2 - \sigma_i^2)$$

It is obvious that $h(\lambda)$ is an increasing function when $\sigma_j > \sigma_i$, and a decreasing function when $\sigma_j < \sigma_i$, for $\lambda > 0$. The following results then can be acquired easily

$$|\alpha u_0/u_1| = \frac{\sigma_1}{\sigma_0}\sqrt{u^2 + v^2}\Big|_{\substack{\sigma_i=\sigma_0\\ \sigma_j=\sigma_1}} > \frac{\sigma_1}{\sigma_0}\sqrt{h(0)}\Big|_{\substack{\sigma_i=\sigma_0\\ \sigma_j=\sigma_1}} = \sqrt{\sigma_1/\sigma_0} \gg 1$$

$$|\beta u_1/u_2| = \frac{\sigma_2}{\sigma_1}\sqrt{u^2 + v^2}\Big|_{\substack{\sigma_i=\sigma_1\\ \sigma_j=\sigma_2}} < \frac{\sigma_2}{\sigma_1}\sqrt{h(0)}\Big|_{\substack{\sigma_i=\sigma_1\\ \sigma_j=\sigma_2}} = \sqrt{\sigma_2/\sigma_1} \ll 1$$

here σ_0 is the conductivity of air and is almost equal to zero.

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透层天线电流积分方程的数值解

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摘要:在已有的三层有耗媒质中的透层天线(BPA)电流积分方程的基础上,给出了电流积分方程中的索末菲积分的低频近似解析公式并给出了其物理解释。利用三项式全域电流基函数,求得了透层天线的输入阻抗。采用静态场近似方法,很方便的求出了透层天线在地面的电场分布。

关键词:透层天线;索末菲积分;分层媒质;输入阻抗;随钻测井系统

中图分类号:TN822^{+}.1　　　　**文献标识码**:A

如 何 学 习 天 线 设 计

天线设计理论晦涩高深，让许多工程师望而却步，然而实际工程或实际工作中在设计天线时却很少用到这些高深晦涩的理论。实际上，我们只需要懂得最基本的天线和射频基础知识，借助于 HFSS、CST 软件或者测试仪器就可以设计出工作性能良好的各类天线。

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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…

详情浏览：http://www.edatop.com/peixun/antenna/116.html

关于易迪拓培训：

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，一直致力和专注于微波、射频、天线设计研发人才的培养；后于 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。

我们的课程优势：

※ 成立于 2004 年，10 多年丰富的行业经验

※ 一直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

※ 视频课程、既能达到了现场培训的效果，又能免除您舟车劳顿的辛苦，学习工作两不误

※ 经验丰富的一线资深工程师主讲，结合实际工程案例，直观、实用、易学

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