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NOTEBOOK

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回顾 天线

一. 引言

辐射：馈线 → 空间

接收：空间 → 馈线

二. 辐射场

1. 电流源和磁流源的矢量位

令 $\vec{H}_e = \frac{1}{\mu}\nabla\times\vec{A}_e$ ①

代入 $\nabla\times\vec{E}_e = -j\omega\mu\vec{H}_e$

得 $\nabla\times(\vec{E}_e + j\omega\vec{A}_e) = 0$

令 $\vec{E}_e + j\omega\vec{A}_e = -\nabla\varphi_e$

代入 $\nabla\times\vec{H}_e = j\omega\varepsilon\vec{E}_e + \vec{J}_e$

得 $\frac{1}{\mu}\nabla\times\nabla\times\vec{A}_e = j\omega\varepsilon(-\nabla\varphi_e - j\omega\vec{A}_e) + \vec{J}_e$

即 $\nabla^2\vec{A}_e + k^2\vec{A}_e = -\mu\vec{J}_e + \nabla(\nabla\cdot\vec{A}_e + j\omega\varepsilon\mu\varphi_e)$

洛伦兹规范 $\nabla\cdot\vec{A}_e + j\omega\varepsilon\mu\varphi_e = 0$

得 $\nabla^2\vec{A}_e + k^2\vec{A}_e = -\mu\vec{J}_e$ ②

且 $\varphi_e = -\frac{1}{j\omega\varepsilon\mu}\nabla\cdot\vec{A}_e$ ③

代入 $\vec{E}_e + j\omega\vec{A}_e = -\nabla\varphi_e$

得 $\vec{E}_e = -j\omega\vec{A}_e - \frac{1}{j\omega\varepsilon\mu}\nabla(\nabla\cdot\vec{A}_e)$ ④

解②得 $\vec{A}_e(\vec{r}) = \frac{\mu}{4\pi}\int\frac{e^{-jk|\vec{r}-\vec{r}'|}}{|\vec{r}-\vec{r}'|}\vec{J}(\vec{r}')\,dv'$ ⑤

将⑤代入①和④可算出由电流源 $\vec{J}_e$ 产生的电场 $\vec{E}_e$ 和 $\vec{H}_e$.

令 $\vec{E}_m = -\frac{1}{\varepsilon}\nabla\times\vec{A}_m$ ⑥

代入 $\nabla\times\vec{H}_m = j\omega\varepsilon\vec{E}_m$

得 $\nabla\times(\vec{H}_m + j\omega\vec{A}_m) = 0$

令 $\vec{H}_m + j\omega\vec{A}_m = -\nabla\varphi_m$

代入 $\nabla\times\vec{E}_m = -j\omega\mu\vec{H}_m - \vec{J}_m$

得 $-\frac{1}{\varepsilon}\nabla\times\nabla\times\vec{A}_m = -j\omega\mu(-\nabla\varphi_m - j\omega\vec{A}_m) - \vec{J}_m$

即 $\nabla^2\vec{A}_m + k^2\vec{A}_m = -\varepsilon\vec{J}_m + \nabla(\nabla\cdot\vec{A}_m + j\omega\varepsilon\mu\varphi_m)$

洛伦兹规范 $\nabla\cdot\vec{A}_m + j\omega\varepsilon\mu\varphi_m = 0$

得 $\nabla^2\vec{A}_m + k^2\vec{A}_m = -\varepsilon\vec{J}_m$ ⑦

且 $\varphi_m = -\frac{1}{j\omega\varepsilon\mu}\nabla\cdot\vec{A}_m$ ⑧

代入 $\vec{H}_m + j\omega\vec{A}_m = -\nabla\varphi_m$

得 $\vec{H}_m = -j\omega\vec{A}_m - \frac{j}{\omega\varepsilon\mu}\nabla(\nabla\cdot\vec{A}_m)$ ⑨

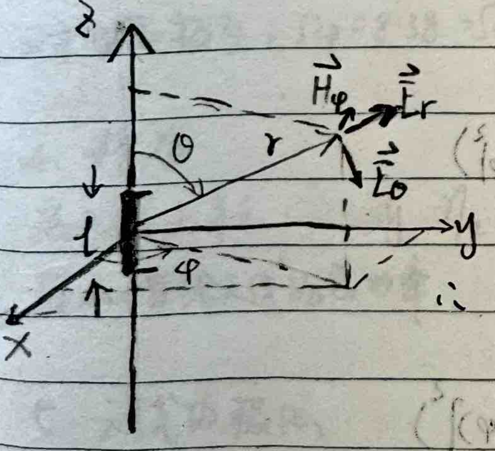
解⑦得 $\vec{A}_m(\vec{r}) = \frac{\varepsilon}{4\pi}\int\frac{e^{-jk|\vec{r}-\vec{r}'|}}{|\vec{r}-\vec{r}'|}\vec{J}_m(\vec{r}')\,dV'$ ⑩

将⑩代入⑥和⑨可算出由磁流源 $\vec{J}_m$ 产生的 $\vec{E}_m$ 和 $\vec{H}_m$.

2. 电偶极子辐射场

电偶极子：长为 l 沿 z 向放置的无限小电流源 I.

$$\vec{A}(r) = \frac{\mu I l e^{-jkr}}{4\pi r}\hat{z} = A_z\hat{z}$$



$$\begin{cases} A_r = A_z\cos\theta = \frac{\mu I l e^{-jkr}}{4\pi r}\cos\theta \\ A_\theta = -A_z\sin\theta = -\frac{\mu I l e^{-jkr}}{4\pi r}\sin\theta \\ A_\varphi = 0 \end{cases}$$

$$\therefore \begin{cases} H_\varphi = \frac{I l e^{-jkr}}{4\pi}\left(\frac{jk}{r} + \frac{1}{r^2}\right)\sin\theta \\ E_r = \frac{I l e^{-jkr}}{4\pi}\left(\frac{2\eta}{r^2} + \frac{2}{j\omega\varepsilon r^3}\right)\cos\theta \\ E_\theta = \frac{I l e^{-jkr}}{4\pi}\left(\frac{j\omega\mu}{r} + \frac{\eta}{r^2} + \frac{1}{j\omega\varepsilon r^3}\right)\sin\theta \\ E_\varphi = H_r = H_\theta = 0 \end{cases}$$

近区场 $kr < 1$

$$\begin{cases} E_r \approx -j\eta\frac{I l e^{-jkr}}{2\pi k r^3}\cos\theta \\ E_\theta \approx -j\eta\frac{I l e^{-jkr}}{4\pi k r^3}\sin\theta \\ H_\varphi \approx \frac{I l e^{-jkr}}{4\pi r^2}\sin\theta \end{cases}$$

远区场 $kr \gg 1$

$$\begin{cases} E_r \approx 0 \\ E_\theta \approx j\eta\frac{k I l e^{-jkr}}{4\pi r}\sin\theta \\ H_\varphi \approx j\frac{k I l e^{-jkr}}{4\pi r}\sin\theta \approx \frac{E_\theta}{\eta} \end{cases}$$

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由 $\vec{S}=\frac{1}{2}(\vec{E}\times\vec{H}^*)=\hat{r}S_r+\hat{\theta}S_\theta+\hat{\varphi}\cdot 0$

得
$$\begin{cases} S_r=\frac{1}{2}E_\theta H_\varphi^*=\frac{\eta}{8}\left(\frac{Il}{\lambda}\right)^2\frac{\sin^2\theta}{r^2}\left[1-j\frac{1}{(kr)^3}\right] \\ S_\theta=-\frac{1}{2}E_r H_\varphi^*=j\eta\frac{kIl^2\cos\theta\sin\theta}{16\pi^2r^3}\left[1+\frac{1}{(kr)^2}\right] \end{cases}$$

总辐射功率
$$P=\oint\vec{S}\cdot d\vec{s}=\int_0^{2\pi}d\varphi\int_0^{\pi}\vec{S}\cdot\hat{r}r^2\sin\theta d\theta$$
$$=\eta\frac{\pi}{3}\left(\frac{Il}{\lambda}\right)^2\left[1-j\frac{1}{(kr)^3}\right]$$

实部为电偶极子辐射功率
$$P_{rad}=\eta\frac{\pi}{3}\left(\frac{Il}{\lambda}\right)^2$$
$$=\frac{1}{2}\left[\eta\frac{2\pi}{3}\left(\frac{l}{\lambda}\right)^2\right]I^2$$
$$=\frac{1}{2}\left[80\pi^2\left(\frac{l}{\lambda}\right)^2\right]I^2$$

三. 天线的基本参数(一)

1. 辐射方向图

① 远区场中 $\vec{E}$、$\vec{H}$ 仅有 $\hat{\theta}$、$\hat{\varphi}$ 分量
$$\vec{E}=\hat{\theta}E_\theta+\hat{\varphi}E_\varphi$$
$$\vec{H}=\frac{1}{\eta}(\hat{\varphi}E_\theta-\hat{\theta}E_\varphi)$$
$$\vec{S}=\frac{1}{2}(\vec{E}\times\vec{H}^*)=\hat{r}\frac{1}{2\eta}(|E_\theta|^2+|E_\varphi|^2)$$

② 远区场中为球面波
$$E_\theta=\frac{f_\theta(\theta,\varphi)}{r}e^{-jkr}$$
$$E_\varphi=\frac{f_\varphi(\theta,\varphi)}{r}e^{-jkr}$$
$$\therefore S_r=\frac{1}{2\eta r^2}(|f_\theta(\theta,\varphi)|^2+|f_\varphi(\theta,\varphi)|^2)$$

功率辐射方向图 $F(\theta,\varphi)=r^2S_r=\frac{1}{2\eta}(|f_\theta(\theta,\varphi)|^2+|f_\varphi(\theta,\varphi)|^2)$

归一化辐射方向图 $F_n(\theta,\varphi)=F(\theta,\varphi)/F_{max}(\theta,\varphi)=S_r/S_{rmax}$ (方向图、波瓣图)

对于电偶极子 $F_n(\theta,\varphi)=\sin^2\theta$ $\begin{cases} E面 & \varphi=const \\ H面 & \theta=90° \end{cases}$

2. 天线立体角

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波瓣主体角 $\Omega_p = \int_{4\pi} F_n(\theta,\varphi)\, d\Omega$

主瓣主体角 $\Omega_M = \int_{主} F_n(\theta,\varphi)\, d\Omega$

旁瓣主体角 $\Omega_m = \int_{旁} F_n(\theta,\varphi)\, d\Omega$

3. 方向性

设 P_T 为辐射功率 $D(\theta,\varphi) = \dfrac{S_r(\theta,\varphi)}{P_T/4\pi r^2}$

若天线无耗 $= \dfrac{4\pi S_r(\theta,\varphi) r^2}{P_T}$

若空间无耗 $= \dfrac{4\pi S_r(\theta,\varphi) r^2}{\int_{4\pi} S_r(\theta,\varphi) r^2 d\Omega}$

$= 4\pi F(\theta,\varphi) / \int_{4\pi} F(\theta,\varphi)\, d\Omega$

$= 4\pi F_n(\theta,\varphi) / \int_{4\pi} F_n(\theta,\varphi)\, d\Omega$

$= 4\pi F_n(\theta,\varphi) / \Omega_p$

$= D_0 F_n(\theta,\varphi)$

其中，最大方向性 $D_0 = \dfrac{4\pi}{\Omega_p} \xlongequal{F_n=1} D(\theta_0,\varphi_0)$

对电偶极子，$\Omega_p = 8.38 = \Omega_M$，$D_0 = 1.5$。

4. 增益

若天线无耗 $\eta_e = \dfrac{P_r}{P_{in}}$

P_e 为馈给天线的总功率.

5. 天线的极化

最大增益方向 辐射波的极化：天线的极化

6. 天线的输入阻抗

目的：天线与馈线 特性阻抗匹配

输入电阻：取决于损耗电阻、辐射电阻

输入电抗：取决于近区场.

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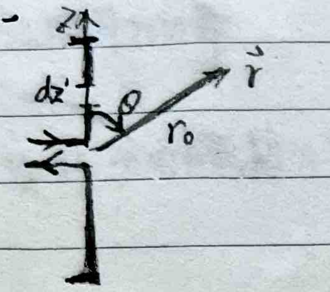
四、略

五、振子天线

1. 对称振子



根据电偶极子的辐射场，远区场

$$dE_\theta \approx j\eta \frac{kI(z')e^{-jkr}}{4\pi r}\sin\theta\, dz'$$

$$\approx j\eta \frac{kI(z')e^{-jk(r_0 - z'\cos\theta)}}{4\pi r_0}\sin\theta\, dz'$$

设 $I(z') \approx \begin{cases} I_0 \sin\left[k\left(\frac{l}{2} - z'\right)\right] & 0 \le z' \le \frac{l}{2} \\ I_0 \sin\left[k\left(\frac{l}{2} + z'\right)\right] & -\frac{l}{2} \le z' < 0 \end{cases}$

得 $$E_\theta = j\eta \frac{ke^{-jkr_0}}{4\pi r_0}\sin\theta I_0 \int_{-\frac{l}{2}}^{0} \sin\left[k\left(\frac{l}{2}+z'\right)\right] e^{jkz'\cos\theta} dz'$$

$$+ j\eta \frac{ke^{-jkr_0}}{4\pi r_0}\sin\theta I_0 \int_0^{\frac{l}{2}} \sin\left[k\left(\frac{l}{2}-z'\right)\right] e^{jkz'\cos\theta} dz'$$

$$= j\eta \frac{I_0 e^{-jkr_0}}{2\pi r_0}\left[\frac{\cos\left(\frac{kl}{2}\cos\theta\right) - \cos\left(\frac{kl}{2}\right)}{\sin\theta}\right]$$

$$H_\varphi = E_\theta / \eta$$

方向图 $$F_n(\theta, \varphi) = \frac{\cos\left(\frac{kl}{2}\cos\theta\right) - \cos\left(\frac{kl}{2}\right)}{\sin\theta}$$

半波振子：$l = \frac{\lambda}{2}$

$$E_\theta = j\eta \frac{I_0 e^{-jkr_0}}{2\pi r_0}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]$$

$$H_\varphi = j \frac{I_0 e^{-jkr_0}}{2\pi r_0}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]$$

$$F_n(\theta, \varphi) = \frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta} \quad \underline{\theta = \frac{\pi}{2}} \quad 1$$

$$S_r = \frac{1}{2\eta}|E_\theta|^2 = \frac{2\eta I_0^2}{(4\pi r_0)^2}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]^2$$

$$P_{rad} = \int_{4\pi} S_r r_0^2 d\Omega \approx 0.609 \frac{\eta I_0^2}{2\pi}$$

$$R_{rad} = \frac{P}{I_0^2/2} = \frac{0.609\eta}{\pi} \approx 73\Omega$$

$$Z_{in} \approx R_{rad} = 73\Omega$$

2. 折叠振子

方向性与半波振子相同。$Z_i = 4Z_{\frac{\lambda}{2}} = 292\Omega$，易与 300Ω 平行双线阻抗匹配。

$R_{rad} \approx 2Z_{\frac{\lambda}{2}} = 146\Omega$。

六、阵列天线

1. 二元阵

$$E_\theta = E_{\theta_1} + E_{\theta_2} = j\frac{\eta k I l}{4\pi}\left[\frac{e^{-j(kr_1 - \frac{\xi}{2})}}{r_1}\sin\theta_1 + \frac{e^{-j(kr_2 + \frac{\xi}{2})}}{r_2}\sin\theta_2\right]$$

$$\approx j\frac{\eta kIl}{4\pi r}e^{-jkr}\sin\theta\left[e^{j(kd\cos\theta+\beta)/2}+e^{-j(kd\cos\theta+\beta)/2}\right]$$

$$=j\frac{\eta kIl}{4\pi r}e^{-jkr}\sin\theta\cdot 2\cdot\cos\frac{kd\cos\theta+\beta}{2}$$

$$=j\frac{\eta kIl}{2\pi r}e^{-jkr}f(\theta\cdot\varphi)$$

2. N元线阵

$$f_a(\psi)=a_0+a_1e^{j\psi}+\cdots+a_{N-1}e^{j(N-1)\psi},\quad \psi=kd\cos\theta+\beta$$

当等幅时 $f_a(\psi)=\frac{\sin N\psi/2}{\sin\psi/2}$

归一化 $f_{an}(\psi)=\frac{1}{N}\frac{\sin N\psi/2}{\sin\psi/2}=Sa(N\psi/2)$.

主瓣最大点: $\psi=0$

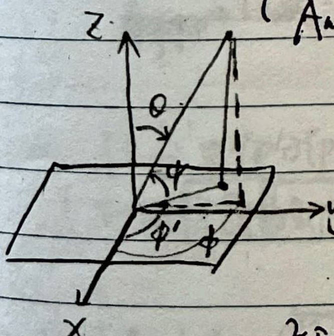
波瓣零点: $\frac{N\psi_0}{2}=p\pi,\ p=\pm1,\pm2,\dots,\pm\frac{N}{2}$

旁瓣最大点: $\frac{N\psi_m}{2}=\frac{2p+1}{2}\pi,\ p=1,\pm2,\dots,-\frac{N}{2}$

七、口面天线:

$$\begin{cases}\text{对相位}: r\approx r_0-r'\cos\xi=r_0-\vec{r}'\cdot\hat{r}\\ \text{对幅度}: r\approx r_0\end{cases}$$

$$\begin{cases}\vec{A}_e=\frac{\mu}{4\pi}\int_S\vec{J}_{es}\frac{e^{-jkr}}{r}ds'\approx\frac{\mu e^{-jkr_0}}{4\pi r_0}\int_S\vec{J}_{es}\cdot e^{jkr'\cos\xi}ds'\\ \vec{A}_m=\frac{\varepsilon}{4\pi}\int_S\vec{J}_{ms}\frac{e^{-jkr}}{r}ds'\approx\frac{\varepsilon e^{-jkr_0}}{4\pi r_0}\int_S\vec{J}_{ms}\cdot e^{jkr'\cos\xi}ds'\end{cases}$$



$$r'\cos\xi=\vec{r}'\cdot\hat{r}=x'\sin\theta\cos\varphi+y'\sin\theta\sin\varphi$$

电导体等效+镜像原理: $\vec{J}_{ms}=-2\hat{n}\times\vec{E}_a,\ \vec{J}_{es}=0$

矩形口面, 均匀场强分布时 $\vec{J}_{ms}=-2\hat{n}\times\vec{E}_a=\hat{x}\cdot2E_0$.

$$\vec{A}_m=\hat{x}\frac{\varepsilon e^{-jkr_0}}{4\pi r_0}\int_{-\frac{b}{2}}^{\frac{b}{2}}\int_{-\frac{a}{2}}^{\frac{a}{2}}2E_0e^{jk(x'\sin\theta\cos\varphi+y'\sin\theta\sin\varphi)}dx'dy'$$

$$=\hat{x}\frac{\varepsilon e^{-jkr_0}}{2\pi r_0}E_0ab\,Sa\left(\frac{ka}{2}\sin\theta\cos\varphi\right)\cdot Sa\left(\frac{kb}{2}\sin\theta\sin\varphi\right).$$

$$\therefore E_\theta\approx-j\omega\eta A_{m\varphi}=j\frac{abkE_0e^{-jkr_0}}{2\pi r_0}\sin\varphi\,Sa\left(\frac{ka}{2}\sin\theta\cos\varphi\right)Sa\left(\frac{kb}{2}\sin\theta\sin\varphi\right)$$

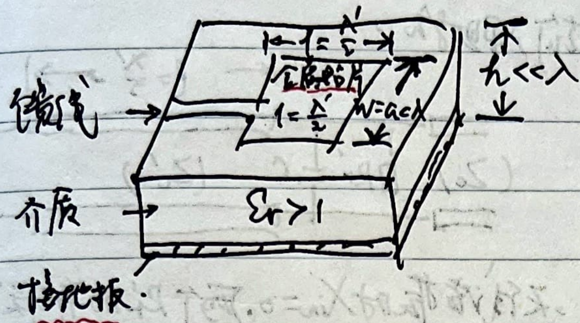
$$E_\varphi\approx-j\omega\eta A_{m\theta}=j\frac{abkE_0e^{-jkr_0}}{2\pi r_0}\cos\theta\cos\varphi\,Sa\left(\frac{ka}{2}\sin\theta\cos\varphi\right)Sa\left(\frac{kb}{2}\sin\theta\sin\varphi\right)$$

$$H_\theta\approx-E_\varphi/\eta$$

$$H_\varphi\approx E_\theta/\eta.$$

§13. 微带天线

1. 矩形微带天线结构



2. 优缺点

a. 具有平面结构，易于与导弹、卫星等飞行体的表面共形.

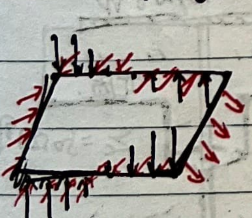
b. 适合于印刷电路技术批量生产

c. 易于构成天线阵.

d. 频带窄，功率容量小，辐射效率低.

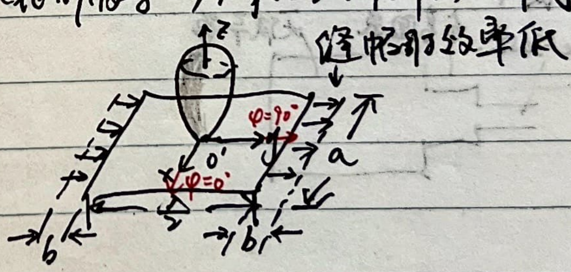
3. 矩形微带天线的分析方法和场方向图函数：传输线模型理论

只需将单缝看作单元天线(矩形缝隙口)，再二元阵列即可



$L=\frac{\lambda_g}{2}$ 电场分布图

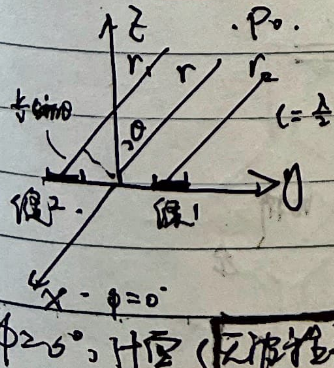
(两端开路 驻波分布)



剩余场分量分布图.

(其余场分量在+z方向的辐射相互抵消)

△ $\varphi=90°$，E面 (有波程差 $\frac{L}{2}\sin\theta$).

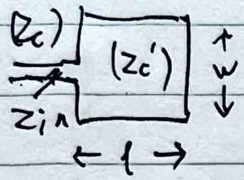


$$f_{E面,总}=f_{E面}(\theta)\cdot f_{a\text{Ⅱ}}=Sa\left(\frac{kb}{2}\sin\theta\right)\cdot 2\cos\left(\frac{kL}{2}\sin\theta\right)$$
$$=Sa\left(\frac{kb}{2}\sin\theta\right)\cdot 2\cos\left(\frac{\pi}{2}\sin\theta\right).$$
$$f_{H面,总}=f_{H面}(\theta)\cdot 2=Sa\left(\frac{ka}{2}\sin\theta\right)\cdot 2\cos\theta.$$

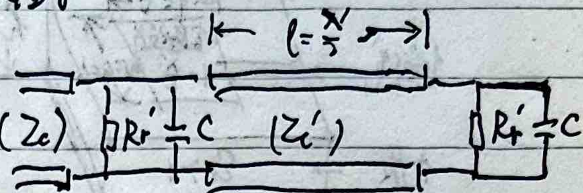
△ $\varphi=0°$，H面 (无波程差 同相叠加).

4. 等效电路和输入阻抗.

A 矩形微带天线的等效电路和其输入阻抗:



等效.



输入阻抗 $Z_{in} = \frac{R_r'}{2}$ (∵ $l=\frac{\lambda'}{2}$, 天线谐振时 $X_{in}=0$. 两个 R_r' 相并联).

研究表明,单缝
$$\begin{cases} R_r' \simeq 120\frac{\lambda}{W}\Omega. & (W \geq 2\lambda) \\ R_r' \simeq 90\left(\frac{\lambda}{W}\right)^2\Omega & (W < 0.35\lambda) \\ R_r' \simeq \frac{1}{W/120\lambda - 1/60\pi^2}\Omega & (0.35\lambda \leq W < 2\lambda) \end{cases}$$

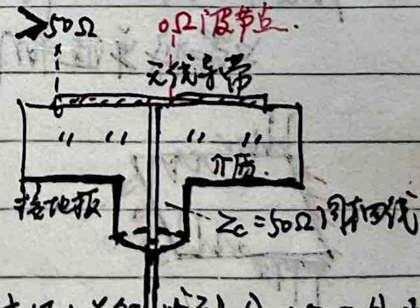
$W\uparrow \to R_r'$

∴ 改变 R_r' 可改 W, 改工作频率可改 l.

5. 馈电方式.

a) 侧馈.

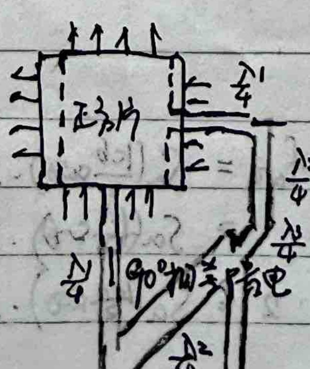
b) 底馈.



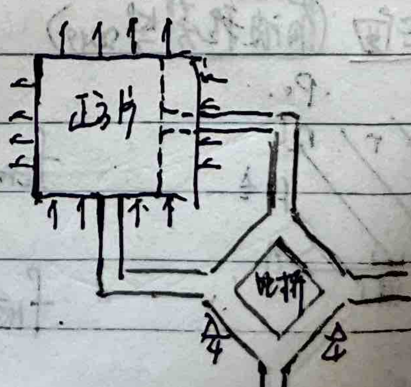
在导体板上选取找到 $R_{in}=50\Omega$ 处,接同轴线

6. 圆极化技术

a) T形分支馈电.



b) 分支线电桥馈电



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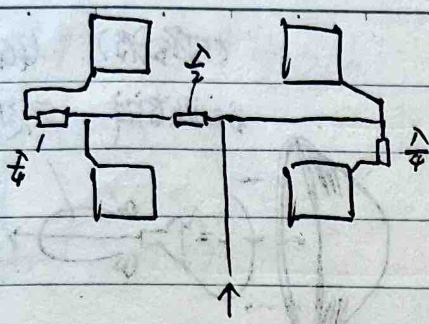
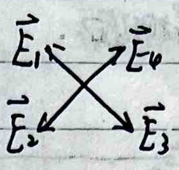
c)



4单元阵列圆极化天线



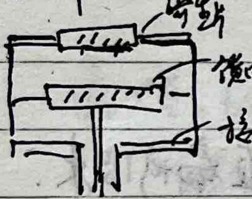
右旋圆极化.

7. 圆极化波的测量方法.

旋转线极化天线测量不同角度的辐射强度.

8. 宽频带、双频微带天线.

叠片展宽频带：

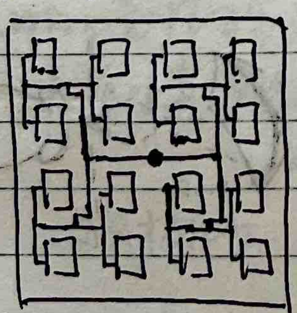


两板的频带相近，展宽频带.

两板的频带相隔较远，则双频带.

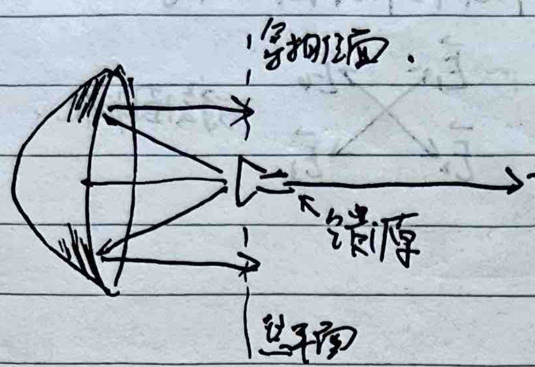
9. 天线阵.

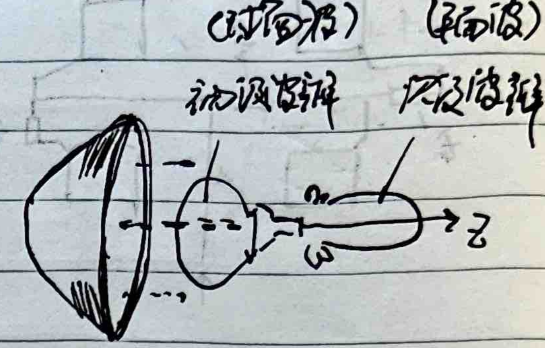
16单元阵列天线，T型分支并联馈电



§14 抛物面天线.

1. 引言.





旋转抛物面的光学性质：(厘米波段、毫米波段适用).

① 反射线平行于轴；平行于轴的光线入射时聚焦于焦点.

② 焦点发射的光线到焦平面的距离相等，为2f.

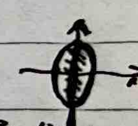
2. 各种形式的抛物面天线.

(a) 针状波束.

跟踪雷达、卫星地面站

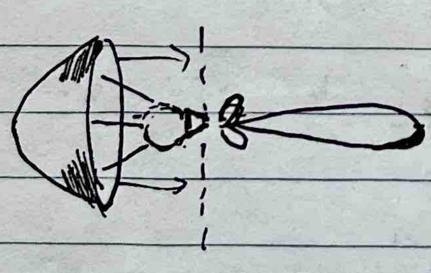
旋转抛物面天线

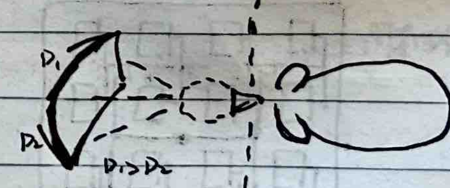
(b) 垂直扇形波束.



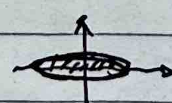
警戒雷达，防颠簸.

切割抛物面(矩形口面).



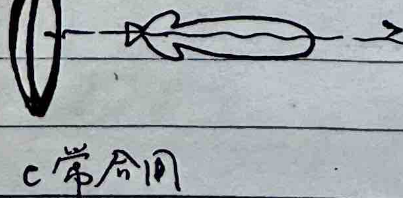


(c) 切割抛物面，



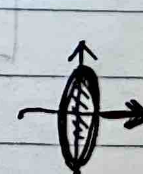
测高雷达：通过测俯仰角.

水平扇形波束(椭圆形口面).



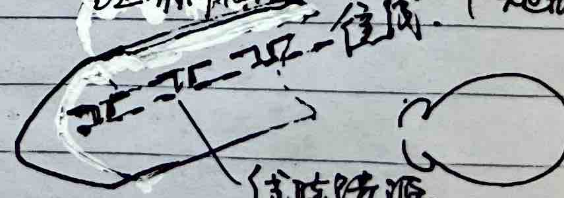
a、c常合用

(d). 抛物柱面



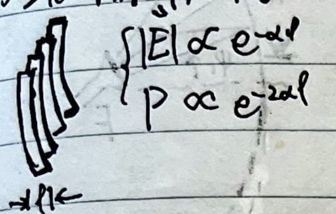
海用导航雷达.

垂直扇形波束 (矩形口面).



3. 减小风阻和重量的方法.

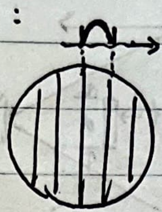
金属栅片结构.



$$\begin{cases}|E|\propto e^{-\alpha l}\\ P\propto e^{-2\alpha l}\end{cases}$$

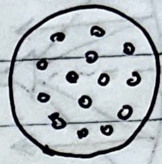
减小风阻.

$\vec{E}$:



垂直极化截止式反射.

波长越短，缝孔越小



$a<\frac{\lambda}{2}\Rightarrow a<\frac{\lambda}{4}\ (\frac{\lambda}{8}\sim\frac{\lambda}{10})$

打孔 $\phi<\frac{\lambda}{4}\ (\frac{\lambda}{8}\sim\frac{\lambda}{10})$

4. 不同焦距的抛物面天线.

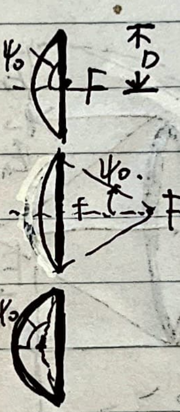
(a) 中焦距 $\psi_0=\frac{\pi}{2}$, $f/D=0.25$.

(b) 长焦距. $\psi_0<\frac{\pi}{2}$, $f/D>0.25$. $(0.25\sim0.6)$.

远旁瓣特性好

(c) 短焦距. $\psi_0>\frac{\pi}{2}$, $f/D<0.25$.

抗干扰特性好.



5. 不同焦距时的口面场分布 (E_s 的分布)

(a) 中焦距. 中心与边缘有交叉极化

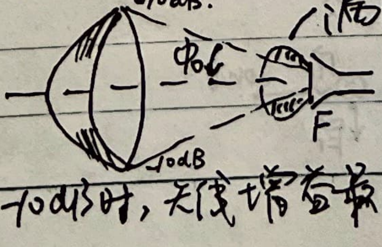
极点：在边缘处

(b) 长焦距 交叉极化小.

(c) 短焦距. 有反向极化，少用.

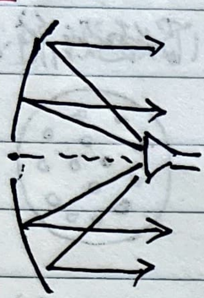
6. 截获功率和漏射功率.

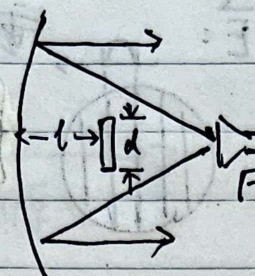
口面截获系数 $\eta_A=P_{rs}$ (截获功率) / P_r (馈源辐射功率).

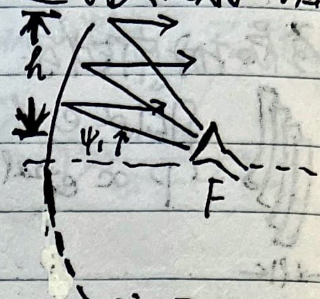


口面边缘场强照射电平为 -10dB 时，天线增益最大.

7. 反射面对馈源的相互影响及消除方法　　问题：遮挡、反射入馈源





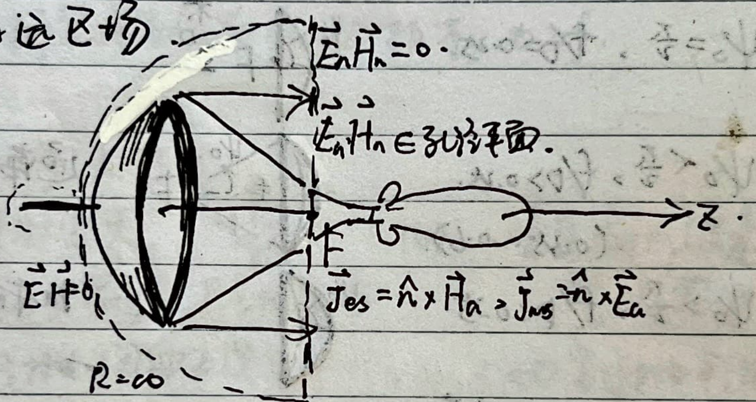


① 抛物面顶点开圆孔.　② 顶点前安装补偿圆盘.　③ 偏置馈源法.

$S=\frac{\lambda^2}{4\pi}G$ 馈源（馈源增益）　$d=\sqrt{\frac{4f\lambda}{\pi}}$，$l=(2n+1)\frac{\lambda}{4}-\frac{\lambda}{4\pi}$.　$\psi_1=a_1(h/f)^2+a_2(h/f)-a_3$ (rad)

8. 求抛物面天线辐射区场

I. 口面场法.



馈源 $\vec E$、$\vec H$
↓
孔径平面 $\vec J_{es}$、$\vec J_{ms}$
↓
$\vec A_m$、$\vec A$
↓
$\vec E$、$\vec H$

a. 求口面上场 $\vec E_a$、$\vec H_a$：由 $\boxed{\text{初级馈源}}$ 和 $\boxed{\text{抛物面反射特性}}$.

b. 孔径平面上 $\vec E_a$、$\vec H_a$，其余为0

c. 由场源等效原理 + 镜像定理，求 $\vec J'_{es}$、$\vec J'_{ms}$

d. 由 $\vec J'_{ms}$ 求 $\vec A_m$，$\vec J'_{es}$ 求 $\vec A$.

e. 由 $\vec A_m$ 或 $\vec A$ 算 $\vec E$ 和 $\vec H$.

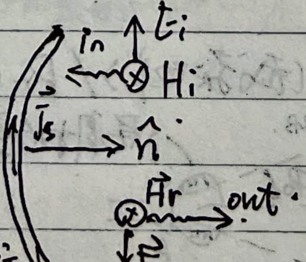
II. 镜面电流法：

a. $\vec J_s=(\hat n\times\vec H_i)+(\hat n\times\vec H_r)$
$=2(\hat n\times\vec H_i)$. 切向H相等

b. 由初级馈源求 $\vec H_i\in S_i$

c. 由 $\vec J_s$ 求 $\vec A$，积分在凹面进行

d. 由 $\vec A$ 求 $\vec E$、$\vec H$



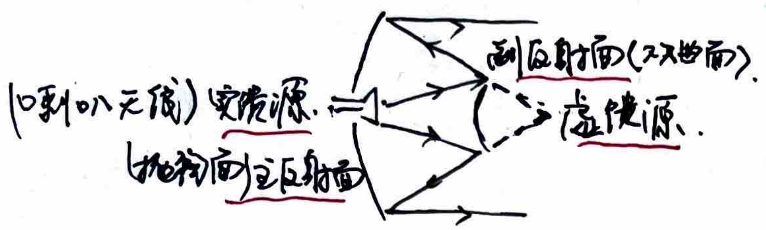
抛物面 $\vec H_i$
↓
抛物面 $\vec J_s$
↓
$\vec A$
↓
$\vec E$、$\vec H$

（科目：　　　）**数　学　作　业　纸**

编号：　　　班级：　　　姓名：　　　第　　页

§15 卡塞格伦天线

①结构及工作原理

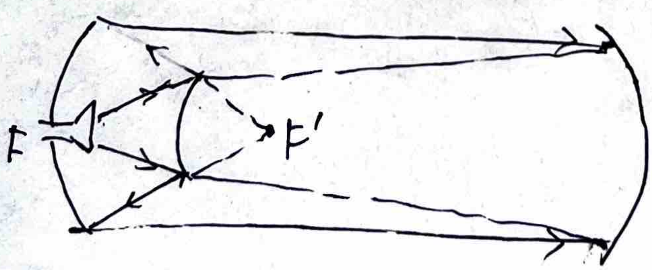


②优点：

1) 口面上场分布容易调到最佳，有两反射面可调节.
2) 结构紧凑、馈线短 —— 后馈
3) 阻抗匹配好 —— 双曲面为扩散反射.
4) "具有长焦距抛物面的电性能"和"机械性能"皆好。

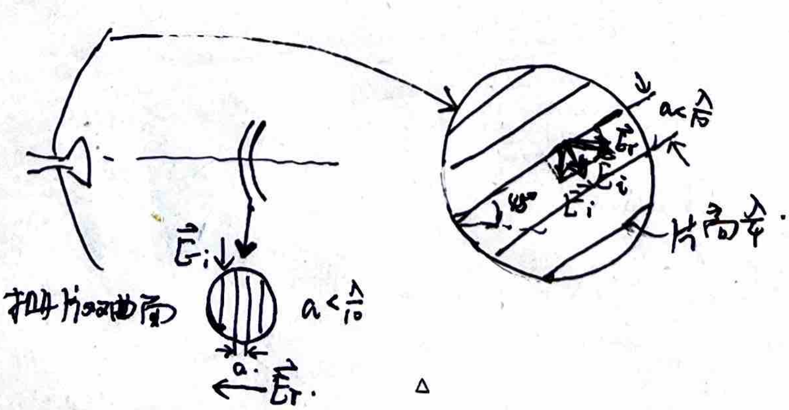
③两种分析方法:

a. 虚馈源法(等效馈源法).
b. 等效抛物面法.

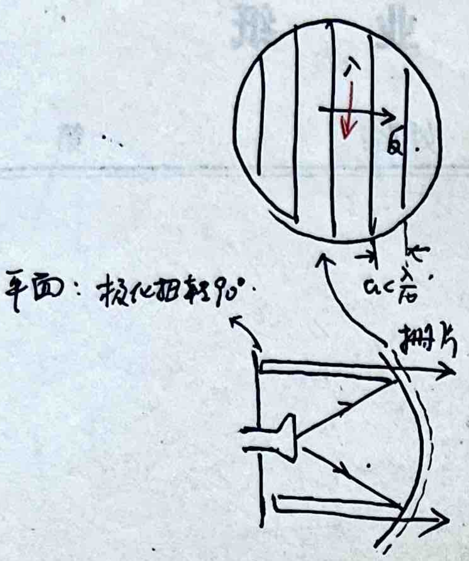


馈源、双曲面、抛物面定后，等效抛物面唯一

④副反射面的遮挡可用"极化扭转90°"来解决.



⑤ 变形卡塞格伦天线.



考 6.22 周三 下午 2:30—4:30

逸馆报告厅。

答疑：伟 1003# 62773748. 6月21日

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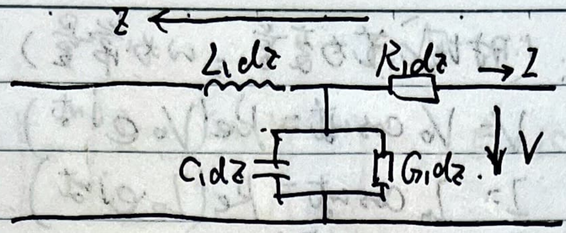
15

初步复习

一. 传输线:



无耗时: $R_1=G_1=0$

频域: 电容 $V(j\omega)=\frac{1}{j\omega C}I(j\omega)$

电感 $V(j\omega)=j\omega L\, I(j\omega)$

电阻 $V(j\omega)=R\, I(j\omega)$.

频域中的无耗传输线方程: $dV(j\omega,z)=I(j\omega,z)j\omega L_1\,dz$

$dI(j\omega,z)=V(j\omega,z)j\omega C_1\,dz$.

解得 $V(j\omega,z)=V_{0+}(j\omega)e^{j\omega\sqrt{L_1C_1}z}+V_{0-}(j\omega)e^{-j\omega\sqrt{L_1C_1}z}$. ①

$I(j\omega,z)=I_{0+}(j\omega)e^{j\omega\sqrt{L_1C_1}z}+I_{0-}(j\omega)e^{-j\omega\sqrt{L_1C_1}z}$. ②

$=\frac{1}{\sqrt{L_1/C_1}}\left[V_{0+}(j\omega)e^{j\omega\sqrt{L_1C_1}z}-V_{0-}(j\omega)e^{-j\omega\sqrt{L_1C_1}z}\right]$

$Z_c\triangleq\frac{1}{\sqrt{L_1/C_1}}$, $\cancel{V_+\triangleq V_{0+}j\omega}$

$V_+(j\omega,z)\triangleq V_{0+}(j\omega)e^{j\omega\sqrt{L_1C_1}z}$, $V_-(j\omega,z)\triangleq V_{0-}(j\omega)e^{-j\omega\sqrt{L_1C_1}z}$.

$I_+(j\omega,z)\triangleq V_+(j\omega,z)/Z_c$, $I_-(j\omega,z)\triangleq -V_-(j\omega,z)/Z_c$.

则 $V(j\omega,z)=V_+(j\omega,z)+V_-(j\omega,z)$ ③

$I(j\omega,z)=I_+(j\omega,z)+I_-(j\omega,z)$ ④

$=V_+(j\omega,z)/Z_c-V_-(j\omega,z)/Z_c$

将①②作反变换, 得 $V(t,z)=V_{0+}(t+\sqrt{L_1C_1}z)+V_{0-}(t-\sqrt{L_1C_1}z)$. ⑤

$I(t,z)=I_{0+}(t+\sqrt{L_1C_1}z)+I_{0-}(t-\sqrt{L_1C_1}z)$. ⑥

边界条件 $\begin{cases}V(j\omega,0)=V_{0+}(j\omega)+V_{0-}(j\omega)=I(j\omega,0)Z_L(j\omega)\\ I(j\omega,0)=\frac{1}{Z_c}[V_{0+}(j\omega)-V_{0-}(j\omega)]\end{cases}$

$\begin{cases}V(j\omega,l)=V_{0+}(j\omega)e^{j\omega\sqrt{L_1C_1}l}+V_{0-}(j\omega)e^{-j\omega\sqrt{L_1C_1}l}=V_g(j\omega)-I(j\omega,l)Z_g(j\omega)\\ I(j\omega,l)=\frac{1}{Z_c}(V_{0+}(j\omega)e^{j\omega\sqrt{L_1C_1}l}-V_{0-}(j\omega)e^{-j\omega\sqrt{L_1C_1}l}).\end{cases}$

消去 $I(j\omega,0)$, $I(j\omega,l)$, 得. $[V_{0+}(j\omega)+V_{0-}(j\omega)]/Z_L(j\omega)=[V_{0+}(j\omega)-V_{0-}(j\omega)]/Z_c$.

$[V_{0+}(j\omega)A+V_{0-}(j\omega)B]/Z_g(j\omega)=V_g(j\omega)/Z_g(j\omega)-[V_{0+}(j\omega)A-V_{0-}(j\omega)B]/Z_c$

可解出 $V_{0+}(j\omega)$ 和 $V_{0-}(j\omega)$

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二. 时谐场下的运算规律：(时域t为变量，ω为常量).

时谐场：$V=V_0\cos\omega t=Re(V_0e^{j\omega t})$，$\tilde{V}\triangleq V_0e^{j\omega t}$

$I=I_0\cos\omega t=Re(I_0e^{j\omega t})$，$\tilde{I}\triangleq I_0e^{j\omega t}$.

电容：$\cancel{I(t)=C\frac{dI(t)}{dt}}$；

$I(t)=C\frac{du(t)}{dt}=j\omega C\cancel{\frac{d}{dt}Re(V_0e^{j\omega t})}$

$=C\,Re(j\omega V_0e^{j\omega t})$

$=\cancel{j\omega CV_0\cdot\omega CV_0e^{j\omega t}}$

$=Re(j\omega C\cdot V_0e^{j\omega t})$

$\tilde{I}(t)=j\omega CV_0e^{j\omega t}=j\omega C\cdot\tilde{V}$ ①

电感：$V(t)=L\frac{dI(t)}{dt}$

$=Re(j\omega L\,I_0e^{j\omega t})$

$\tilde{V}(t)=j\omega L\cdot\tilde{I}$ ②

电阻：$V(t)=R\,I(t)$

$\tilde{V}(t)=R\,\tilde{I}(t)$. ③

∵ $e^{j\omega t}$因子对线性运算的不变性，∴ $\tilde{V}(\vec{x},t)$、$\tilde{I}(\vec{x},t)$可直接写成$V(t)$ $I(t)$

∵ ①②③式，∴ 可将电容电感电阻等阻抗引入传输线.

上述规律是"路"的规律，"场"的规律类似，将$\tilde{E}(\vec{x},t)$，$\tilde{H}(\vec{x},t)$直接写成$E(\vec{x})$、$H(\vec{x})$

三. 微波网络的研究方法.

为了定义任意截面沿z方向单模传输的均匀波导参考面上的模式电压和模式电流，作如下规定：只考虑沿一个方向的传播 $\hat{z}$ or $-\hat{z}$.

(1) $V_t(z)\propto E_T$，$I_t(z)\propto H_T$.

(2) $\tilde{P}=V_t(z)\cdot I_t^*(z)$

(3) $Z_c=V_t(z)/I_t(z)$.

进入网络：内向波 $a=V_+/\sqrt{Z_c}$ 向网络移：$a'=ae^{j\theta}$，$b'=be^{j\theta}$

离开网络：外向波 $b=V_-/\sqrt{Z_c}$. 向外侧移：$b'=be^{-j\theta}$；$a'=ae^{j\theta}$.

对N端口网络，定义

$$\begin{pmatrix} b_1 \\ \vdots \\ b_n \end{pmatrix} = \begin{pmatrix} S_{11} & \cdots & S_{1n} \\ & \ddots & \\ S_{n1} & \cdots & S_{nn} \end{pmatrix} \begin{pmatrix} a_1 \\ \vdots \\ a_n \end{pmatrix}, \quad \vec{b} = [S]\vec{a}.$$

$\left\{\begin{array}{l} i\text{口匹配：} S_{ii}=0 \\ i\text{口接匹配负载：} \Gamma_i'=0. \end{array}\right.$

(1) 互易：$S_{ij}=S_{ji}$，$[S]=[S]^T$.

(2) 无耗：$[S]^H\cdot[S]=I$

对二端口网络

(1) 互易：$S_{12}=S_{21}$

(2) 无耗：$|S_{12}|=|S_{21}|$，$|S_{11}|=|S_{22}|$，$|S_{11}|^2+|S_{21}|^2=1$，$|S_{12}|^2+|S_{22}|^2=1$.

1) 端口1匹配则端口2匹配，电压传输系数的模为1.

$|S_{11}|=|S_{22}|=0$，$|S_{12}|=|S_{21}|=1$.

2) 不能实现单向隔离，隔离是双方向的

$|S_{12}|=|S_{21}|=0$.

3) 双向隔离时，产生全反射

$|S_{12}|=|S_{21}|=0$，$|S_{11}|=|S_{22}|=1$.

(3) 对称：$S_{11}=S_{22}$，$S_{12}=S_{21}$.

1) 本征方程

$$\begin{bmatrix} S_{11}-S_j & S_{12} \\ S_{12} & S_{11}-S_j \end{bmatrix} \begin{bmatrix} u_1^j \\ u_2^j \end{bmatrix} = 0.$$

(2) 本征值：

$$\left\{\begin{array}{l} S_1 = S_{11}+S_{12} \\ S_2 = S_{11}-S_{12}. \end{array}\right. \Rightarrow \left\{\begin{array}{l} S_{11} = \dfrac{S_1+S_2}{2} \\ S_{12} = \dfrac{S_1-S_2}{2}. \end{array}\right.$$

对本征矢：$S_1 \Rightarrow u_1 = \frac{1}{\sqrt{2}}(1,1)^T$

$S_2 \Rightarrow u_2 = \frac{1}{\sqrt{2}}(1,-1)^T$.

意义：以 u_1 作为两个端口的入射波 a，则 $b=[S]u_1=S_1u_1$，反射系数为 S_1.

以 u_2 作为两个端口的入射波 a，则 $b=[S]u_2=S_2u_2$ 反射系数为 S_2.

因 u_1 的两个分量对两个端口来说等幅同向，对应于中间开路情况

因 u_2 的两个分量对两个端口来说等幅反向，对应于中间短路情况.

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(4) 指标.

1) 电压传输系数 $T_v = S_{21}$

2) 功率传输系数 $T_p = |S_{21}|^2$.

3) 插入衰减 $L = \frac{1}{|S_{21}|^2}$

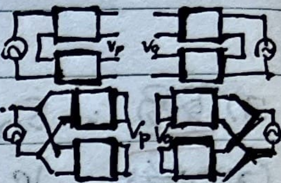
4) 插入相移 $\theta_{21} = \arg S_{21}$.

5) 群时延 $t_d = \frac{d}{d\omega}\theta_{21}$.

6) 输入驻波比 $\rho = \frac{1+|\Gamma_1|}{1-|\Gamma_1|} = \frac{1+|S_{11}|}{1-|S_{11}|}$.

(5) 阻抗矩阵: 串联: $[Z] = \sum_i [Z_i]$ $V_p = V_q = 0$: 有效串联

导纳矩阵: 并联 $[Y] = \sum_i [Y_i]$. $V_p = V_q = 0$: 有效并联.



级联矩阵: 级联: $[A] = \prod_i [A_i]$. (用 V_1 I_1 V_2 $-I_2$ 表示).

传输矩阵: 级联: $[t] = \prod_i [t_i]$ (用 a_1 b_1 a_2 b_2 表示)

串并矩阵: 串并联: $[H] = \sum [H_i]$

并串矩阵: 并串联: $[G] = \sum [G_i]$

四. 波导到传输线. (时谐、均匀、无源、无耗)

1. 纵横关系. (算子: $\nabla = \nabla_T + \nabla_z$; $\nabla^2 = \cdots = \nabla_T^2 + \nabla_z^2$).

无源、时谐时:

沿 $\hat{z}$ 传 (-偏z ≙ k_z)

$$(k^2-k_z^2)\vec{H}_T = -j\omega\varepsilon\hat{z}\times\nabla_T E_z - jk_z\nabla_T H_z.$$

$$(k^2-k_z^2)\vec{E}_T = j\omega\mu\hat{z}\times\nabla_T H_z - jk_z\nabla_T E_z.$$

沿 $-\hat{z}$ 传: (偏z ≙ $-k_z$).

$$(k^2-k_z^2)\vec{H}_T = j\omega\varepsilon\hat{z}\times\nabla_T E_z + jk_z\nabla_T H_z$$

$$(k^2-k_z^2)\vec{E}_T = -j\omega\mu\hat{z}\times\nabla_T H_z + jk_z\nabla_T E_z.$$

2. TE波: $E_z = 0$.

沿 $\hat{z}$ 传:

$$\vec{H}_T = -\frac{jk_z}{k^2-k_z^2}\nabla_T H_z.$$

$$\vec{E}_T = \frac{j\omega\mu}{k^2-k_z^2}\hat{z}\times\nabla_T H_z.$$

$$\vec{E}_T = -\frac{\omega\mu}{k_z}\hat{z}\times\vec{H}_T \triangleq -\eta_{TE}\hat{z}\times\vec{H}_T$$

沿 $-\hat{z}$ 传 反号.

$$\vec{E}_T = \eta_{TE}\hat{z}\times\vec{H}_T$$

边界条件: $\left.\frac{\partial H_z}{\partial n}\right|_{边界} = 0$.

算子：$\nabla=\nabla_T+\nabla_z$，$\nabla^2=\cdots=\nabla_T^2+\nabla_z^2$　　T：横，z：纵

$$\nabla^2 H_z+k^2H_z=0$$

沿 $\hat z$ 传：$H_z=H_z(T)\cdot e^{-jk_z z}$.　　沿 $-\hat z$ 传：$H_z=H_z(T)\cdot e^{jk_z z}$.

都得到 $\nabla_T^2 H_z(T)+(k^2-k_z^2)H_z(T)=0$

1. TE波：$E_z=0$.

$$\begin{cases}\nabla^2H_z+k^2H_z=0\longrightarrow \nabla_T^2H_z(T)+(k^2-k_z^2)H_z(T)=0,\ H_z(z)=e^{-jk_z z}\ \text{or}\ e^{jk_z z}\\ \left.\dfrac{\partial H_z}{\partial n}\right|_{\text{边界}}=0\end{cases}$$

解得沿 $\hat z$ 和 $-\hat z$ 传的 H_z；k_z。定义 $\eta_{TE}(\Omega)=\frac{\omega\mu}{k_z}$.

2. TM波：$H_z=0$

$$\begin{cases}\nabla^2E_z+k^2E_z=0\longrightarrow \nabla_T^2E_z(T)+(k^2-k_z^2)E_z(T)=0,\ E_z(z)=e^{-jk_z z}\ \text{or}\ e^{jk_z z}\\ E_z|_{\text{边界}}=0.\end{cases}$$

解得沿 $\hat z$ 和 $-\hat z$ 传的 E_z；k_z。定义 $\eta_{TM}(\Omega)=\frac{k_z}{\omega\varepsilon}$

3. TEM波：$E_z=0$，$H_z=0$，$\vec W_T\triangleq\vec E_T$ or $\vec H_T$

$$\begin{cases}\nabla^2\vec W_T+k^2\vec W_T=0\longrightarrow \nabla_T^2\vec W_T(T)=0\to\nabla_T\cdot\vec W_T=\nabla_T\times\vec W_T=0,\ W_T(z)=e^{-jk_z z}\ \text{or}\ e^{jk_z z}\\ \text{无}\left(\int_a^b E\cdot dl=V\right)\searrow \text{单根空心金属波导中不存在TEM波.}\end{cases}$$

解得沿 $\hat z$ 和 $-\hat z$ 传的 E_x、E_y、H_x、H_y。定义 $\eta_{TEM}=\sqrt{\frac{\mu}{\varepsilon}}$，$k_z=k$.

4. 由无源、时谐和 $H(z)$、$E(z)=e^{-jk_z z}$ or $e^{jk_z z}$，得纵横关系式 ~~(……)~~.

沿 $\hat z$：

$$\begin{cases}(k^2-k_z^2)\vec H_T=-j\omega\varepsilon\hat z\times\nabla_T E_z-jk_z\nabla_T H_z\\ (k^2-k_z^2)\vec E_T=j\omega\mu\hat z\times\nabla_T H_z-jk_z\nabla_T E_z\end{cases}$$

沿 $-\hat z$：

$$\begin{cases}(k^2-k_z^2)\vec H_T=j\omega\varepsilon\hat z\times\nabla_T E_z+jk_z\nabla_T H_z\\ (k^2-k_z^2)\vec E_T=-j\omega\mu\hat z\times\nabla_T H_T+jk_z\nabla_T E_z.\end{cases}$$

由此解出TE和TM波的 $\vec H_T$、$\vec E_T$，(根据 H_z、E_z.)

5. 参数.

(1) 导波波数：$k_z\geq0$.（$v_{p}=\frac{\omega}{k_z}$，$v_g=\frac{d\omega}{dk_z}$）；导波波长：$\lambda_g=\frac{2\pi}{k_z}$；导波频率：同变量.

(2) 截止波数：$k_c\triangleq\sqrt{k^2-k_z^2}$　　；截止波长：$\lambda_c=\frac{2\pi}{k_c}$；截止频率：$f_c=\frac{k_c\cdot c}{2\pi}$

(1) 是导波的实际参数，有物理意义　　(2) 中参数与波导尺寸有关，用于区分尺寸.

正向：$e^{-jk_z z}$　反向：$e^{jk_z z}$

波腹：$\varphi_F - 2k_z z = 2k\pi$，波节：$\varphi_F - 2k_z z = \pi + 2k\pi$.

6. 传输线描述的依据.

根据1、2、3、各模式具有正交性，对于每一模式，由分离变量的结果（波动方程）：

$$\begin{cases}\frac{\partial^2}{\partial z^2}\vec{E}_T + k_z^2\vec{E}_T = 0 \\ \frac{\partial^2}{\partial z^2}\vec{H}_T + k_z^2\vec{H}_T = 0\end{cases} \Rightarrow \begin{cases}\vec{E}_{T\pm}(T,z) = \vec{E}_{T\pm}(T)\cdot e^{\pm jk_z z}, & \vec{E}_T = \vec{E}_{T+} + \vec{E}_{T-} \\ \vec{H}_{T\pm}(T,z) = \vec{H}_{T\pm}(T)\cdot e^{\pm jk_z z}, & \vec{H}_T = \vec{H}_{T+} + \vec{H}_{T-}\end{cases}$$

$\because$
$$\begin{cases}\vec{E}_T \triangleq V\hat{e}(x,y) \\ \vec{H}_T \triangleq I\hat{h}(x,y)\end{cases} \quad \begin{cases}\vec{E}_{T\pm} \triangleq V_\pm \hat{e}(x,y) \\ \vec{H}_{T\pm} \triangleq I_\pm \hat{h}(x,y)\end{cases} \quad \begin{cases}V \triangleq \int_l \vec{E}\cdot d\vec{l} \\ I \triangleq \oint_c \vec{H}\cdot d\vec{l}\end{cases}$$

$\therefore$
$$\begin{cases}\frac{\partial^2}{\partial z^2}V + k_z^2 V = 0 \\ \frac{\partial^2}{\partial z^2}I + k_z^2 I = 0.\end{cases}$$

解得
$$\begin{cases}V = V_+ + V_- = V_{0+}e^{jk_z z} + V_{0-}e^{-jk_z z} & ① \\ I = I_+ + I_- = I_{0+}e^{jk_z z} + I_{0-}e^{-jk_z z} & ②\end{cases}$$

易证 $V_+/I_+ = -V_-/I_-$

$\therefore$ (1) $Z_c \triangleq V_+/I_+ = -V_-/I_-$，$Y_c = 1/Z_c$.　e.g. 同轴线 $Z_c = 60\sqrt{\frac{\mu_r}{\varepsilon_r}}\ln\frac{b}{a}$

(2) $\Gamma \triangleq V_-/V_+ = -I_-/I_+ = \frac{V_{0-}e^{-jk_z z}}{V_{0+}e^{jk_z z}} = \Gamma_F e^{-j2k_z z} = |\Gamma_F|e^{j(\varphi_F - 2k_z z)}$ ③

$\because \Gamma_F = \frac{Z_F - Z_c}{Z_F + Z_c} = \frac{R_F + jX_F - Z_c}{R_F + jX_F + Z_c}$

$\therefore |\Gamma_F| = \sqrt{\frac{(R_F - Z_c)^2 + X_F^2}{(R_F + Z_c)^2 + X_F^2}}$

$\because$ 无源负载 $R_F \geq 0$

$\therefore |\Gamma_F| \leq 1$

$\therefore |\Gamma| \leq 1$.

(3) $\bar{Z}_i \triangleq \frac{V}{I}/Z_c = \frac{V_+ + V_-}{I_+ + I_-}/Z_c = \frac{V_+(1+\Gamma)}{I_+(1-\Gamma)}/Z_c = \frac{1+\Gamma}{1-\Gamma} = \frac{\bar{Z}_F + j\tan k_z z}{1 + j\bar{Z}_F\tan k_z z}$ ④

(4) $\bar{y}_i \triangleq \frac{I}{V}/Y_c = \cdots = \frac{1-\Gamma}{1+\Gamma} = \frac{\bar{y}_F + j\tan k_z z}{1 + j\bar{y}_F\tan k_z z}$ ⑤

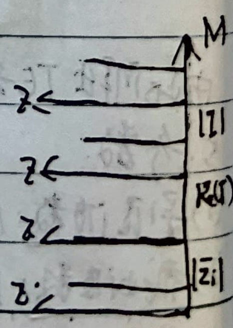
(5) $\rho \triangleq \frac{V_{max}}{V_{min}} = \frac{1+|\Gamma|}{1-|\Gamma|} \in [1,\infty)$

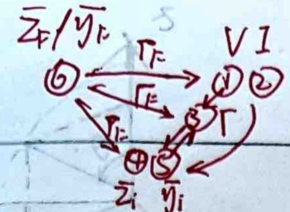
(6) $l_{min} \triangleq \frac{\varphi_F - \pi}{2k_z}$

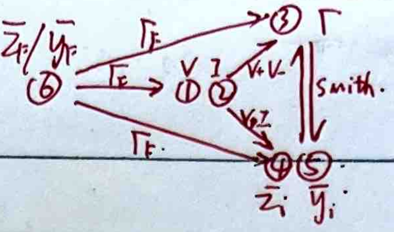
$\therefore \Gamma = \frac{\bar{Z}_i - 1}{\bar{Z}_i + 1} = \frac{1 - \bar{y}_i}{1 + \bar{y}_i}$ ⑥

① 行波状态. $\bar{Z}_F \triangleq \bar{y}_F \triangleq 1$，$\Gamma_F = 0$

$$\begin{cases}V(z) = V_+ = V_{0+}e^{jk_z z}, & |V(z)| = |V_{0+}| \\ I(z) = I_+ = I_{0+}e^{jk_z z}, & |I(z)| = |I_{0+}| \\ \Gamma(z) = 0 \\ \bar{Z}_i(z) = \bar{y}_i(z) = 1.\end{cases}$$

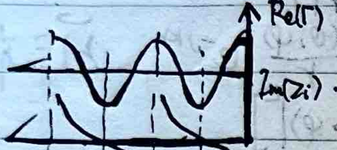




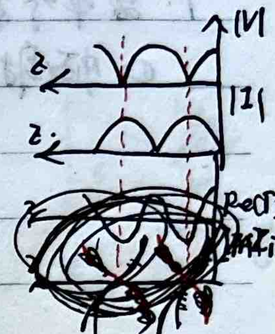


② 纯驻波状态.

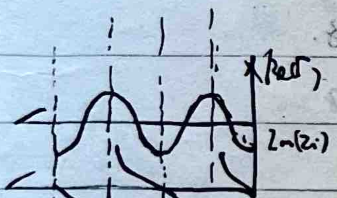
1) 终端开路. $\bar z_F \triangleq \infty / \bar y_F \triangleq 0$, $\Gamma_F = +1$.



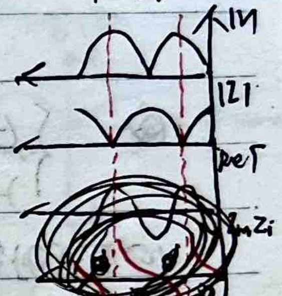
$$\begin{cases} V(z) = 2V_{0+}\cos k_z z, \quad |V(z)| = 2|V_{0+}||\cos k_z z|. \\ I(z) = 2j\dfrac{V_{0+}}{Z_c}\sin k_z z, \quad |I(z)| = 2\left|\dfrac{V_{0+}}{Z_c}\right||\sin k_z z| \\ \Gamma(z) = +e^{-j2k_z z}. \\ \bar z_i(z) = -j\cot k_z z, \quad \bar y_i(z) = j\tan k_z z. \end{cases}$$



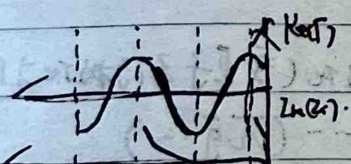
2) 终端短路 $\bar z_F \triangleq 0 / \bar y_F \triangleq \infty$, $\Gamma_F = -1$



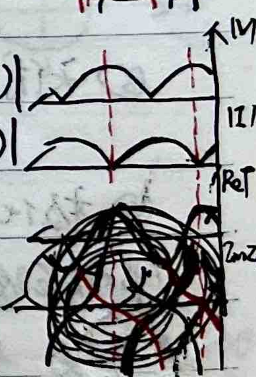
$$\begin{cases} V(z) = j2V_{0+}\sin k_z z, \quad |V(z)| = 2|V_{0+}||\sin k_z z| \\ I(z) = \dfrac{2V_{0+}}{Z_c}\cos k_z z, \quad |I(z)| = 2\left|\dfrac{V_{0+}}{Z_c}\right||\cos k_z z| \\ \Gamma(z) = -e^{-j2k_z z}. \\ \bar z_i(z) = j\tan k_z z, \quad \bar y_i(z) = -j\cot k_z z. \end{cases}$$



3) 终端纯电抗 $\bar z_F \triangleq jx / \bar y_F \triangleq jb$, $|\Gamma_F| = 1$. ($\Gamma_F = e^{j\phi_F}$).

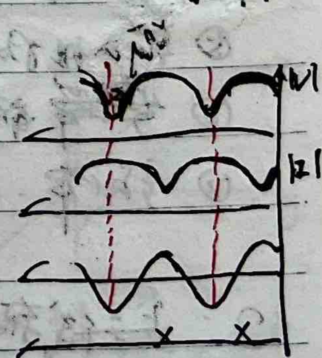


$$\begin{cases} V(z) = 2V_{0+}e^{j\frac{\phi_F}{2}}\cos\left(k_z z - \frac{\phi_F}{2}\right), \quad |V(z)| = 2|V_{0+}|\left|\cos\left(k_z z - \frac{\phi_F}{2}\right)\right| \\ I(z) = 2j\dfrac{V_{0+}}{Z_c}e^{j\frac{\phi_F}{2}}\sin\left(k_z z - \frac{\phi_F}{2}\right), \quad |I(z)| = 2\left|\dfrac{V_{0+}}{Z_c}\right|\left|\sin\left(k_z z - \frac{\phi_F}{2}\right)\right| \\ \Gamma(z) = e^{j(\phi_F - 2k_z z)}. \\ \bar z_i(z) = -j\cot\left(k_z z - \frac{\phi_F}{2}\right), \quad \bar y_i(z) = j\tan\left(k_z z - \frac{\phi_F}{2}\right) \end{cases}$$



③ 驻波状态

$$\begin{cases} V(z) = V_{0+}\left(1 + \Gamma_F e^{-j2k_z z}\right), \quad |V(z)| = |V_{0+}|\left|1 + |\Gamma_F| e^{j(\phi_F - 2k_z z)}\right| \\ I(z) = \dfrac{V_{0+}}{Z_c}\left(1 - \Gamma_F e^{-j2k_z z}\right), \quad |I(z)| = \left|\dfrac{V_{0+}}{Z_c}\right|\left|1 - |\Gamma_F| e^{j(\phi_F - 2k_z z)}\right| \\ \Gamma(z) = |\Gamma_F| e^{j(\phi_F - 2k_z z)} \\ \bar z_i(z) = \dfrac{1 + \Gamma_F e^{-j2k_z z}}{1 - \Gamma_F e^{-j2k_z z}}, \quad \bar y_i(z) = \dfrac{1 - \Gamma_F e^{-j2k_z z}}{1 + \Gamma_F e^{-j2k_z z}}. \end{cases}$$



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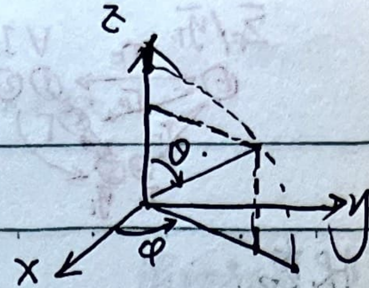


五. 天线

1. 基本参数:

① 辐射方向图，远区场中只有 E·H θ和φ分量，球面波.

$$\vec S = \frac{1}{2}(\vec E\times\vec H^*) = \hat r\,\frac{1}{2\eta}\left(|E_\theta|^2+|E_\varphi|^2\right)$$

$$E_\theta = \frac{f_\theta(\theta,\varphi)}{r}e^{-jkr},\quad E_\varphi = \frac{f_\varphi(\theta,\varphi)}{r}e^{-jkr},\quad \vec H = \frac{1}{\eta}\hat r\times\vec E$$

$$\therefore \vec S = \frac{1}{2\eta r^2}\left[|f_\theta(\theta,\varphi)|^2+|f_\varphi(\theta,\varphi)|^2\right]$$

$$F(\theta,\varphi)\triangleq r^2 S = \frac{1}{2\eta}\left[|f_\theta(\theta,\varphi)|^2+|f_\varphi(\theta,\varphi)|^2\right]$$

$$F_n(\theta,\varphi)\triangleq F(\theta,\varphi)/F_{max}(\theta,\varphi).$$

{ E面方向图：$\vec E$所在平面F, e.g. $\varphi = const$ for 振子.
H面方向图：$\vec H$所在平面F, e.g. $\theta = 90°$ for 振子

② 方向性系数（辐射功率相同时）

$$D(\theta,\varphi)\triangleq \frac{S_r(\theta,\varphi)}{P_r/4\pi r^2}$$

③ 天线增益（输入功率相同时）

$$G(\theta,\varphi)\triangleq \frac{S_r(\theta,\varphi)}{P_{in}/4\pi r^2}$$

④ 极化（最大增益方向）
线极化：平行极化（平行于入射面）、垂直极化（垂直于平行极化方向）.
圆极化：右旋圆极化（右手螺旋法则）、左——（左手——）
椭圆极化：= 两正交振幅不等的圆极化波叠加.

⑤ 输入阻抗 Z_{in} = 馈点电压/馈点电流.（馈电面上）

⑥ 辐射电阻 $R_{rad} = 2P_r/I_0^2$.

⑦ 频带. 满足方向系数、增益等要求的频率范围.

2. ~~天线~~辐射场

电流源 $\vec A_e(\vec r) = \frac{\mu}{4\pi}\int \frac{e^{-jk|\vec r-\vec r'|}}{|\vec r-\vec r'|}\vec J(\vec r')dV'$；磁流源：$\vec A_m(\vec r) = \frac{\varepsilon}{4\pi}\int \frac{e^{-jk|\vec r-\vec r'|}}{|\vec r-\vec r'|}\vec J_m(\vec r')dV'$

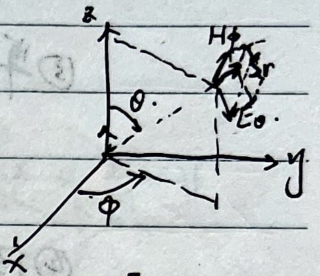
$\vec H_e = \frac{1}{\mu}\nabla\times\vec A_e$ $\qquad$ $\vec H_m = -j\omega\vec A_m - \frac{1}{\omega\varepsilon\mu}\nabla(\nabla\cdot\vec A_m)$

$\vec E_e = -j\omega\vec A_e - \frac{1}{\omega\varepsilon\mu}\nabla(\nabla\cdot\vec A_e).$ $\qquad$ $\vec E_m = -\frac{1}{\varepsilon}\nabla\times\vec A_m$

3. 点天线（$I=I_0e^{j\omega t}$）

近区：$\vec{S}$ 纯虚数，储能场，无辐射.

远区：$\vec{S}$ 实数，辐射场，球面波.



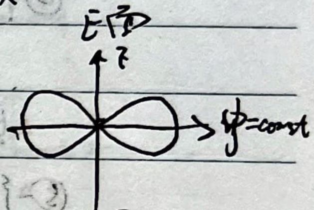
$$H_\phi \approx \frac{I_0 l}{4\pi} e^{-jkr}\frac{jk}{r}\sin\theta = E_\theta/\eta$$

$$E_\theta \approx \frac{I_0 l}{4\pi} e^{-jkr} j\frac{\omega\mu}{r}\sin\theta$$

$$S_r = \frac{1}{2}(\vec{E}_\theta\times\vec{H}_\phi^*)\cdot\hat{r} = \frac{\eta}{8}\left(\frac{I_0 l}{\lambda}\right)^2\frac{\sin^2\theta}{r^2}$$

$$P_r = \oint_S S_r ds = 40\pi^2\left(\frac{I_0 l}{\lambda}\right)^2$$

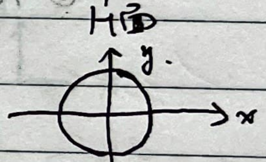
$$R_r = 2P_r/I_0^2 = 80\pi^2\left(\frac{l}{\lambda}\right)^2$$



$f_n(\theta)=\sin\theta$，$F_n(\theta)=\sin^2\theta$

$f_n(\theta)\triangleq\frac{1}{\sqrt{2}} \Rightarrow |\theta_2-\theta_1| = 90°$ 半功率角.

$$D(\theta,\phi)=\frac{S_r}{P_r/4\pi r^2}=\frac{3}{2}\sin^2\theta \xrightarrow{\theta=\frac{\pi}{2}} 1.5 = 1.76dB.$$

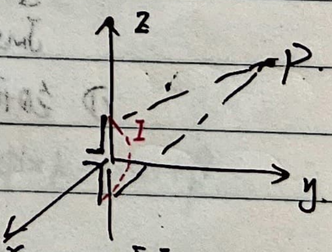


4. 线天线

远区：$dH_\phi = j\frac{I dz}{4\pi}e^{-jkr}\frac{jk}{r}\sin\theta = dE_\theta/\eta$.

$$dE_\theta = \frac{I dz}{4\pi} e^{-jkr} j\frac{\omega\mu}{r}\sin\theta.$$

① 对称振子. $\left(\begin{cases} I(z')=I_0\sin k(\frac{l}{2}-z') & 0\le z'\le\frac{l}{2} \\ I(z')=I_0\sin k(\frac{l}{2}+z') & -\frac{l}{2}\le z'\le 0\end{cases}\right)$ $\begin{cases} \text{算幅度} & R\approx r \\ \text{算相位} & R\approx r-z'\cos\theta\end{cases}$



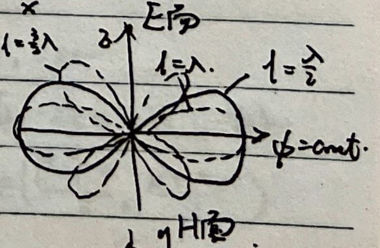
$$H_\phi = j\frac{I_0 e^{-jkr}}{2\pi r}\cdot\frac{\cos(\frac{kl}{2}\cos\theta)-\cos\frac{kl}{2}}{\sin\theta} = E_\theta/\eta$$

$$E_\theta = j\eta\frac{I_0 e^{-jkr}}{2\pi r}\cdot\frac{\cos(\frac{kl}{2}\cos\theta)-\cos\frac{kl}{2}}{\sin\theta}$$

$$S_r \propto \eta\frac{I_0^2}{8\pi^2 r^2}\left[\frac{\cos(\frac{kl}{2}\cos\theta)-\cos\frac{kl}{2}}{\sin\theta}\right]^2$$

$l=\frac{\lambda}{2}$：$P_r = 36.54 I_0^2$

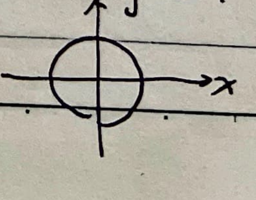
$\Downarrow$ $R_r = 73\Omega = Z_{in}$.



$f_n(\theta) = \frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}$，$F_n(\theta) = \frac{f_n^2(\theta)}{(1-\cos\frac{kl}{2})^2}$

$f_n(\theta)\triangleq\frac{1}{\sqrt{2}} \Rightarrow |\theta_2-\theta_1| = 78°$ 半功率角.

$D(\theta,\phi) \xrightarrow{\theta=\frac{\pi}{2}} 1.64 = 2.15dB$

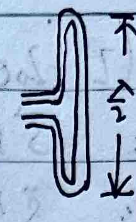


$l=\lambda$：$R_r = 200\Omega$；$l=\frac{3}{2}\lambda$：$R_r=106\Omega$.

No.

Date

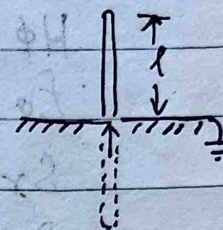
② 折叠振子



$$R_r \approx 2\times 73\,\Omega$$

$$Z_{in} \approx 4\times 73\,\Omega.$$

③ 单极天线

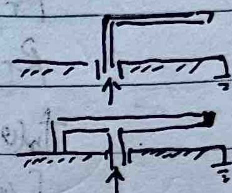


$D = 2D_{对称振子}$.

$l=\frac{\lambda}{4}$时，$D = 2\times 1.64 = 3.4 = 5.16\,dB$.

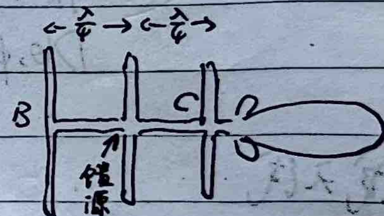
④ 倒L/倒F天线：弯曲的单极天线.

倒F中增加的一段用来增加 R_{rad}.



△ 缝隙天线 —①②③④：终端开路.

⑤△ 行波天线 — 八木.宇田.



1号稍长于$\frac{\lambda}{2}$：Z_{inB} 感性

2号等于$\frac{\lambda}{2}$：在1、3号感应出场($\approx$远场)

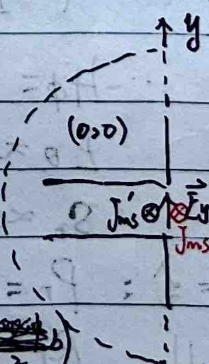
3号稍短于$\frac{\lambda}{2}$：Z_{inC} 容性

∴ 1号、2号或3号、2号在y同相叠加，在$-y$反相叠加。 ①反 ②振 ③引

5. 面天线

远区：电导体/导磁体等效 + 镜像原理.

$\vec{J}_{ms} = -\hat{n}\times\vec{E}$ / $\vec{J}_{es} = \hat{n}\times\vec{H}$.



① 矩形波导口. ($\vec{E}_y = \hat{y}E_0$)　幅：$R\approx r$

$\vec{J}'_{ms} = \hat{x}J'_m = \hat{x}\cdot 2E_0$　相：$R\approx r - r'\cos\psi$

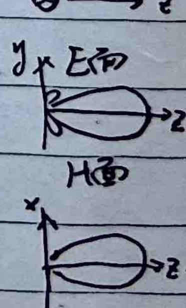
$$\cancel{H_{m\theta} = -j\omega\frac{\varepsilon ab E_0 e^{-jkr}}{2\pi r}\cos\theta\cos\phi\, Sa\left(\frac{k\sin\theta\cos\phi}{2}a\right)Sa\left(\frac{k\sin\theta\sin\phi}{2}b\right)}$$

$$\vec{A}_m = \hat{x}\cdot\frac{\varepsilon e^{-jkr}}{2\pi r}E_0\cdot ab\, Sa\left(\frac{k\sin\theta\cos\phi}{2}a\right)Sa\left(\frac{k\sin\theta\sin\phi}{2}b\right)$$

$A_{mr} = \sin\theta\cos\phi\, A_m$，$A_{m\theta} = \cos\theta\cos\phi\, A_m$，$A_{m\phi} = -\sin\phi\, A_m$

$H_{m\theta}\approx -j\omega A_{m\theta}$，$H_{m\phi}\approx -j\omega A_{m\phi}$

$E_{m\theta}\approx \eta H_{m\phi}$，$E_{m\phi}\approx -\eta H_{m\theta}$



$\phi=90°$: $f_{n,E} = Sa\left(\frac{kb}{2}\sin\theta\right) \triangleq \frac{1}{\sqrt2}$, $2\theta_{0.5,E} \simeq 51°\cdot\frac{\lambda}{b}$.

$\phi=0°$: $f_{n,H} = Sa\left(\frac{ka}{2}\sin\theta\right) \triangleq \frac{1}{\sqrt2}$, $2\theta_{0.5,H} \simeq 51°\frac{\lambda}{a}$.

③ 矩形波导口 ($\vec{E}_y = \hat{y}E_0\sin\frac{\pi}{a}x$)

$$2\theta_{0.5,E} = 68°\frac{\lambda}{b}$$

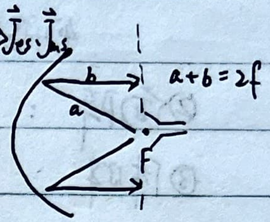
$$2\theta_{0.5,H} = 68°\frac{\lambda}{a}.$$

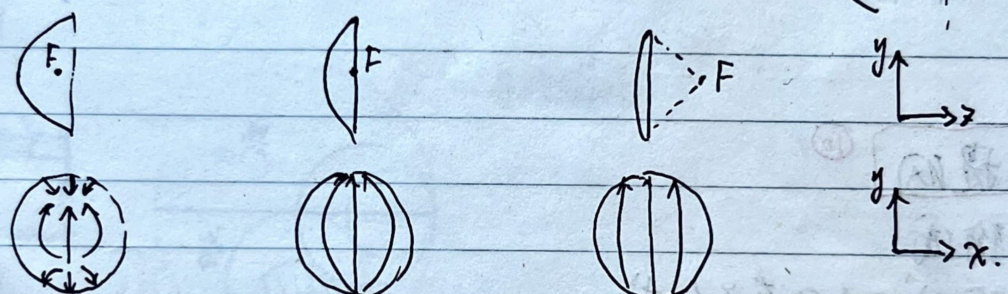
⑤ 抛物面天线.

{ 口面场法: 类似于矩形波导口, 求口面上 $E\cdot H \Rightarrow J_s, J_{ms}$

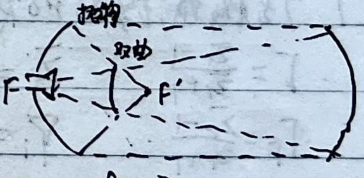
  镜面电流法: 由抛物面上的 J_s 算 $\vec{A}$

短焦距: 反相抵消　中焦距: 交叉极化　长焦距: ✓





④ 卡塞格伦天线: 因"极化扭转 p_0" 削减副反射面遮挡 (波程差 $\lambda/2$).



6. 天线阵 { 虚馈源法 / 等效抛物面法: 见右图.

远区: $f(\theta,\phi) = f_a(\theta,\phi)\cdot f_e(\theta,\phi) \approx f_a(\theta,\phi)$　{ 幅: $R\simeq r$; 相: $R\simeq r\pm\frac{d}{2}\cos\theta$.

阵列因子　单元因子

① 线阵.

1) 二元: $E_\theta = j\frac{k\eta I_0 l}{4\pi}\frac{e^{-jkr}}{r}\sin\theta\cdot 2\cos\frac{kd\cos\theta+\beta}{2}$

$= j\frac{k\eta I_0 l}{4\pi}\frac{e^{-jkr}}{r}f_e(\theta,\phi)\cdot f_a(\theta,\phi)$.

2) N元: $f_{an} = \sin N\frac{\psi}{2}\Big/N\sin\frac{\psi}{2}$　N可整数, $\psi = kd\cos\theta+\beta$.

$\theta_{max} = \cos^{-1}\left(-\frac{\beta}{kd}\right)$, $\psi=0$

$\theta_0 = \cos^{-1}\left(\frac{2p\pi}{Nkd}-\frac{\beta}{kd}\right)$, $\psi=\frac{2p\pi}{N}$, $p=\pm1,\pm2\cdots$

$\theta_{max,副} = \cos^{-1}\left(\frac{2(p+\frac12)\pi}{Nkd}-\frac{\beta}{kd}\right)$, $\psi=\frac{2(p+\frac12)\pi}{N}$, $p=(1,-2),(2,-3),(3,-4)\cdots$

可见区: $\theta\in(0,180°)$, $\psi\in(\beta-kd,\beta+kd)$, $\frac{\psi}{2}\in\left(\frac{\beta-kd}{2},\frac{\beta+kd}{2}\right)$

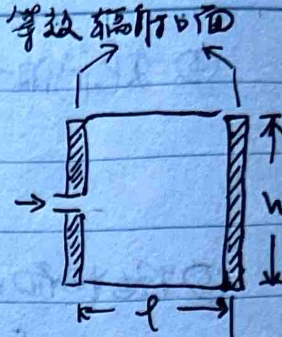
3) 微带天线: 二元面阵列.

$f_{E,a} = Sa\left(\frac{kb}{2}\sin\theta\right)\cdot 2\cos\left(\frac{kd}{2}\sin\theta\right)$

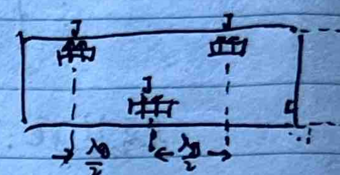
$f_{H,a} = Sa\left(\frac{ka}{2}\sin\theta\right)\cdot 2\cos\theta$.

$$Z_{in} = \frac{R_r}{2}$$

$$R_r \approx \begin{cases} 120\dfrac{\lambda}{W} & W \geq 2\lambda \\ 90\left(\dfrac{\lambda}{W}\right)^2 & W < 0.35\lambda \\ \left[W/120\lambda - 1/60\pi^2\right]^{-1} & 0.35\lambda \leq W < 2\lambda \end{cases}$$



4) 波导开缝天线：切割横向的传输电流，右图：



② 面阵：$f_a = f_{ax} \cdot f_{ay}$

③ 体阵：$f_a = f_{ax} \cdot f_{ay} \cdot f_{az}$.

复习提纲 ⑩

一、传输线

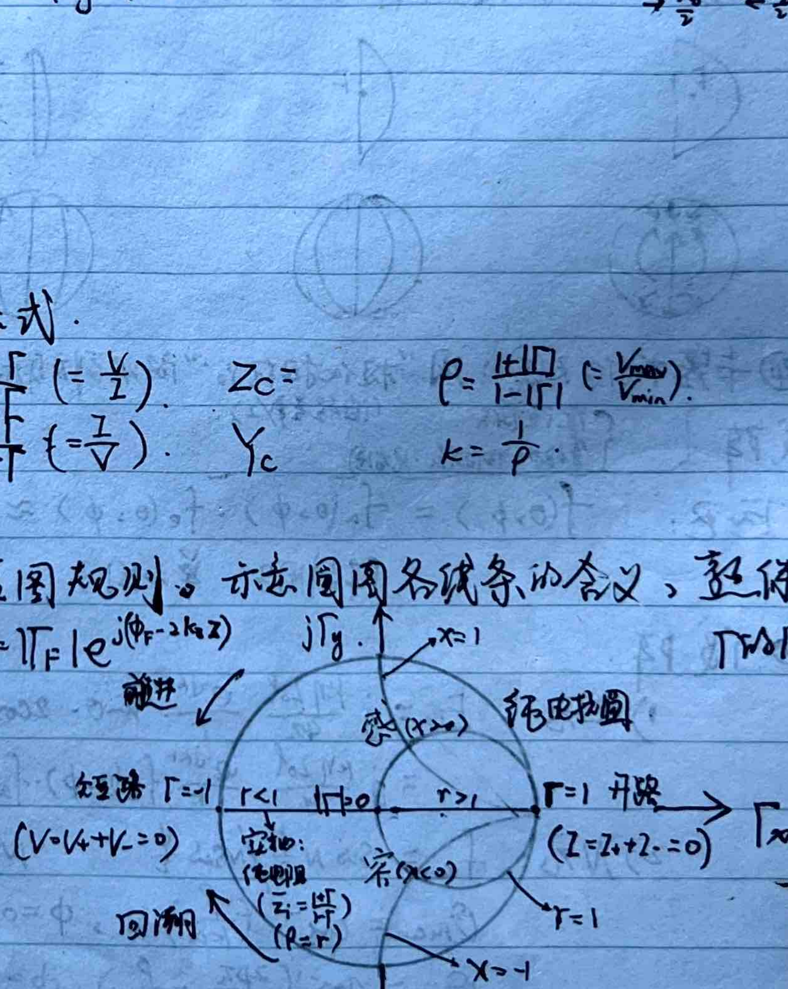
1. 传输线理论的有关公式.

$\Gamma = \frac{\bar{z}_i - 1}{\bar{z}_i + 1}$ $\quad \bar{z}_i = \frac{1+\Gamma}{1-\Gamma} \left(= \frac{V}{I}\right)$. $\quad Z_c =$ $\quad \rho = \frac{1+|\Gamma|}{1-|\Gamma|} \left(= \frac{V_{max}}{V_{min}}\right)$. $\quad \bar{z} = Z_{in}/Z_c$

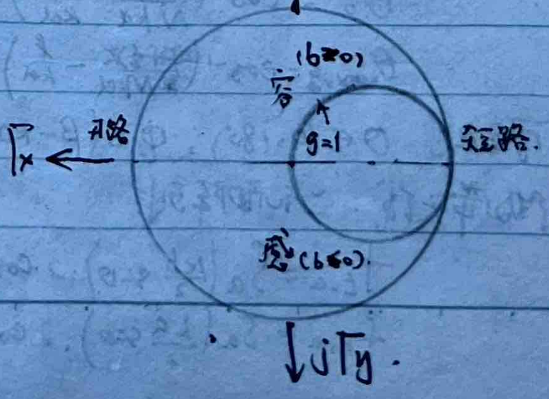
$\Gamma_F = \frac{\bar{z}_F - 1}{\bar{z}_F + 1}$. $\quad \bar{y}_i = \frac{1-\Gamma}{1+\Gamma} \left(= \frac{I}{V}\right)$. $\quad Y_c$ $\quad k = \frac{1}{\rho}$. $\quad \bar{y} = \frac{1}{\bar{z}}$.

2. Smith圆图的构造，查图规则。示意圆图各线条的含义，熟练解题.

阻抗圆图： $\Gamma = |\Gamma_F| e^{j(\phi_F - 2k_0 z)}$ $\quad$ Γ的周期：$\frac{\lambda}{2}$

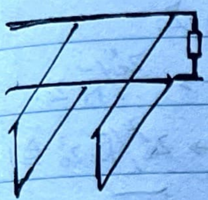


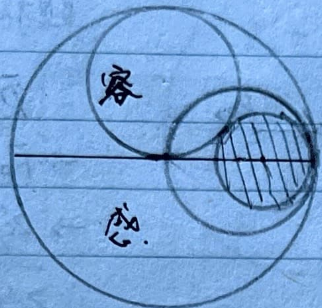
导纳圆图：



①求解盲区

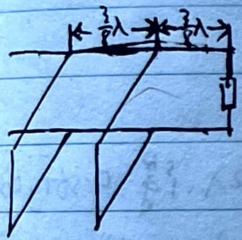
P49 例2.9.（双线调配器的盲区）.

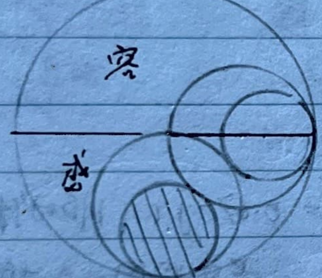




导纳圆图.

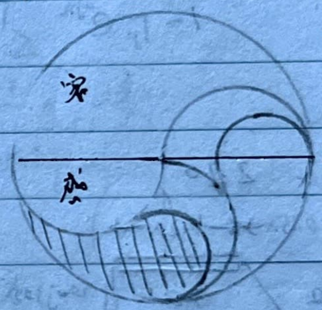
P76. 2.11





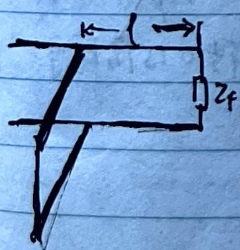
导纳圆图.

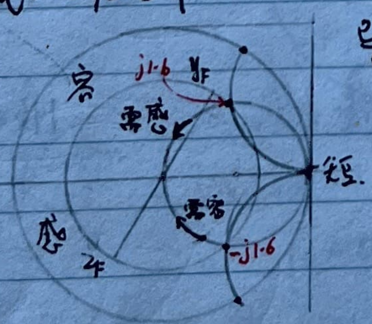
不超全.



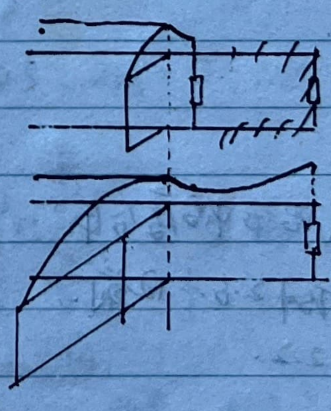
②单线调配器.

P46 例2.6.（线上电压分布）.

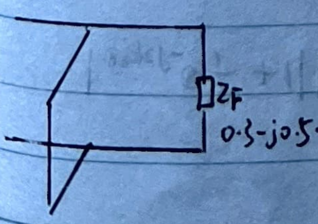


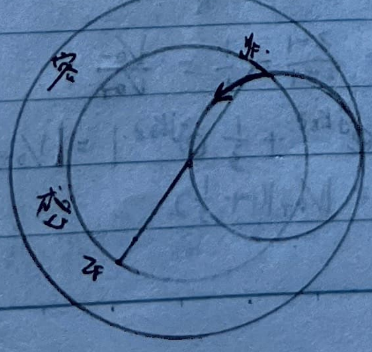


导纳圆图



P75. 2.5.





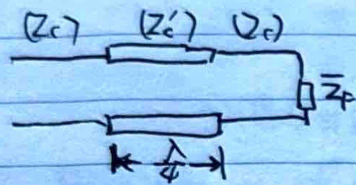
导纳圆图

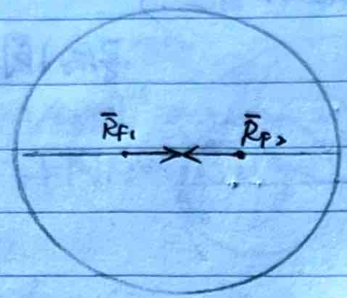
No.

Date

3. 查圆图例题

P48. 例2.8 ($\frac{\lambda}{4}$ 变阻器).





阻抗圆图

$\bar{R}_F \to \bar{R}_F \cdot Z_c$

$\to \bar{R}_F \cdot Z_c / Z_c' \to Z_c' / \bar{R}_F \cdot Z_c$

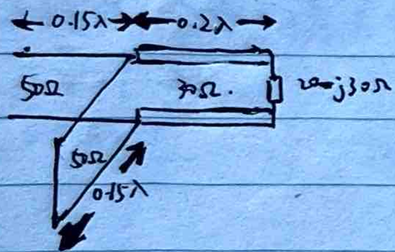
$\to Z_c'^2 / \bar{R}_F Z_c \to Z_c'^2 / \bar{R}_F \cdot Z_c^2 \triangleq 1$

$\therefore Z_c' = \sqrt{\bar{R}_F Z_c^2} = \sqrt{R_F Z_c}$.

$$Z_i = \frac{1+\Gamma}{1-\Gamma} = \frac{1+\Gamma_F e^{-j2k_z z}}{1-\Gamma_F e^{-j2k_z z}}$$

$$= \frac{1+\Gamma_F \cdot e^{-j\pi}}{1-\Gamma_F e^{-j\pi}} = \frac{1-\Gamma_F}{1+\Gamma_F} = \frac{1}{z_F}.$$

P76. 2.10.



$\bar{Z}_F = \frac{20-j30}{30} = 0.67-j1$，顺时针转0.2λ、得. 0.56+j0.31

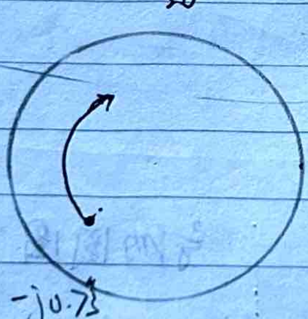
导纳圆图. 按30反归一化，得 10.8+j10.2

按50归一化，得0.216+j0.204.

并联后：~~0.216+j0.204−j0.73~~

$= 0.216 - j0.526$

顺时针转0.15后即可.

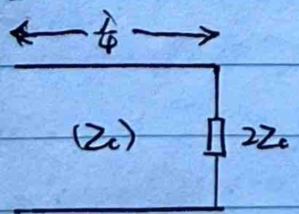


4. 画沿线电压幅度分布.

P40 例2.6. 见前.

P75 2.2.

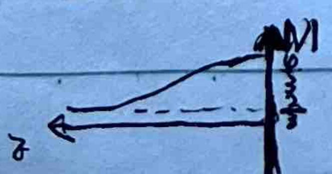
(a).



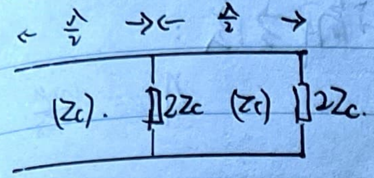
$\Gamma_F = \frac{z_F-1}{z_F+1} = \frac{2-1}{2+1} = \frac{1}{3} = \frac{V_{0-}}{V_{0+}}$

$\therefore |V| = |V_{0+}|\,|e^{jk_z z} + \frac{1}{3} e^{-jk_z z}| = |V_{0+}|\,|1+\frac{1}{3} e^{-j2k_z z}|$

$z=0$ 时，$|V| = |V_{0+}|(1+\frac{1}{3})$.



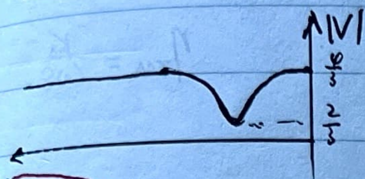
(b)



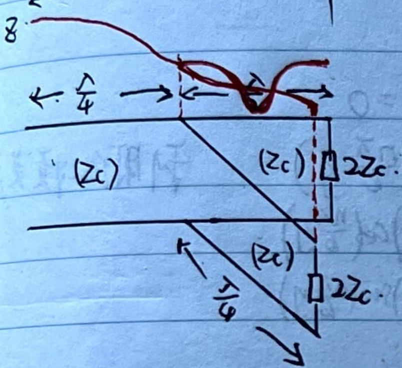
$\Gamma_F = \frac{2-1}{2+1} = \frac{1}{3}$ $\quad |V| = |V_{0+}|\,|1 + \frac{1}{3} e^{-j2k_z z}|$.

$Z'_{左} = \frac{1-\Gamma}{1+\Gamma} = \frac{1-\Gamma_F e^{-j2k_z\cdot\frac{\lambda}{2}}}{1+\Gamma_F e^{-j2k_z\cdot\frac{\lambda}{2}}} = \frac{1-\Gamma_F}{1+\Gamma_F} = 2Z_c$.

$\therefore Z_{中} = 2Z_c // 2Z_c = Z_c$. $\quad \therefore \Gamma_{中} = 0$.



(c).



$\Gamma_{F上} = \frac{1}{3}$, $\Gamma_{F下} = \frac{1}{3}$.

$Z_{中} = 2Z_c // \frac{1}{2}Z_c = \frac{2}{5}Z_c$

$\therefore \Gamma_{中} = \cancel{|V_{0+}|\,|1+\ldots|} \frac{\frac{2}{5}-1}{\frac{2}{5}+1} = -\frac{3}{7}$.

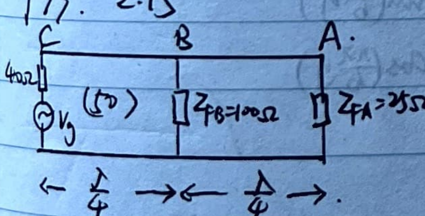
$\therefore |V| = |V_{0+}|\,|1 - \frac{3}{7} e^{-j2k_z z}|$.

5. 入射波电压，反射波电压，总电压，反射系数的概念.

$$V = V_+ + V_-$$

$$\Gamma = \frac{V_-}{V_+}$$

例. 2.15



解：$\bar{Z}_{FA} = 0.5$，$\bar{Z}_{FB} = 2$，$\therefore \bar{Z}_{中} = 2 // 2 = 1$.

$\therefore \bar{Z}_{左} = 1$，$Z_{左} = 50\Omega$.

$\therefore \Gamma_C = 0$，$\dot{V}_B = \dot{V}'_g e^{-jk_z\cdot\frac{\lambda}{4}} = -j\dot{V}'_g$，$P_{FB} = \frac{V_g'^2}{2R_{FB}}$.

又 $\bar{Z}_{FA中} = 2$. $\therefore \Gamma_B = \frac{2-1}{2+1} = \frac{1}{3}$

$\therefore \dot{V}_- = \frac{3}{4}\dot{V}_B$，$\dot{V}_+ = \frac{1}{4}\dot{V}_B$

$\therefore \dot{V}_A = \dot{V}_- e^{-jk_z\frac{\lambda}{4}} + \dot{V}_+ e^{jk_z\frac{\lambda}{4}} = (-j\frac{3}{4} + j\frac{1}{4})\dot{V}_B$

$= -\frac{j}{2}\dot{V}_B = -\frac{1}{2}\dot{V}'_g$，$P_{FA} = \frac{V_g'^2}{8R_{FA}}$

反射系数的方向性！瞬时性！

No.
Date

二. 波导

标量波动方程 $\xrightarrow{\text{边界条件}}$ 纵向分量 $\xrightarrow{\text{纵横关系}}$ 横向分量

1. 熟悉矩形波导、圆波导、同轴线中场型的求解方法：

矩形、圆：

TE: $\begin{cases} \nabla_T^2 H_z(x,y) + k_c^2 H_z(x,y) = 0 \\ \frac{\partial H_z(x,y)}{\partial n}\Big|_{\text{边界}} = 0 \end{cases}$ $\eta_{TE} = \frac{\omega\mu}{k_z}$

TM: $\begin{cases} \nabla_T^2 E_z(x,y) + k_c^2 E_z(x,y) = 0. \\ E_z(x,y)\big|_{\text{边界}} = 0. \end{cases}$ $\eta_{TM} = \frac{k_z}{\omega\varepsilon}.$

矩形TE:

$$\begin{cases} \frac{\partial^2}{\partial x^2} H_z + \frac{\partial^2}{\partial y^2} H_z + k_c^2 H_z = 0. \\ \frac{\partial H_z}{\partial x}\Big|_{x=0,a} = 0, \quad \frac{\partial H_z}{\partial y}\Big|_{y=0,b} = 0 \end{cases}$$

$\therefore H_z(x,y) = H_{mn}\cos\left(\frac{m\pi}{a}x\right)\cos\left(\frac{n\pi}{b}y\right)$. 利用纵横关系得

$$H_x(x,y) = \frac{jk_z}{k_c^2}\left(\frac{m\pi}{a}\right)H_{mn}\sin\left(\frac{m\pi}{a}x\right)\cos\left(\frac{n\pi}{b}y\right).$$

$$H_y(x,y) = \frac{jk_z}{k_c^2}\left(\frac{n\pi}{b}\right)H_{mn}\cos\left(\frac{m\pi}{a}x\right)\sin\left(\frac{n\pi}{b}y\right).$$

$$E_x(x,y) = \eta_{TE} H_y(x,y)$$

$$E_y(x,y) = -\eta_{TE} H_x(x,y).$$

矩形TM:

$$\begin{cases} \frac{\partial^2}{\partial x^2} E_z + \frac{\partial^2}{\partial y^2} E_z + k_c^2 E_z = 0. \\ E_z\big|_{x=0,a} = 0, \quad E_z\big|_{y=0,b} = 0. \end{cases}$$

$\therefore E_z(x,y) = E_{mn}\sin\left(\frac{m\pi}{a}x\right)\sin\left(\frac{n\pi}{b}y\right)$. 利用纵横关系得.

$$E_x(x,y) = -j\frac{k_z}{k_c^2}\left(\frac{m\pi}{a}\right)E_{mn}\cos\left(\frac{m\pi}{a}x\right)\sin\left(\frac{n\pi}{b}y\right).$$

$$E_y(x,y) = -j\frac{k_z}{k_c^2}\left(\frac{n\pi}{b}\right)E_{mn}\sin\left(\frac{m\pi}{a}x\right)\cos\left(\frac{n\pi}{b}y\right)$$

$$H_x(x,y) = -\frac{1}{\eta_{TM}} E_y(x,y)$$

$$H_y(x,y) = \frac{1}{\eta_{TM}} E_x(x,y).$$

圆TE:

$$\begin{cases} \frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial H_z}{\partial r}\right) + \frac{1}{r^2}\frac{\partial^2 H_z}{\partial \varphi^2} + k_c^2 H_z = 0. \\ \frac{\partial H_z}{\partial r}\Big|_{r=R} = 0. \end{cases}$$

$H_z(r,\varphi) = H_{ni} J_n(k_c r)\cos(n\varphi)$. 利用纵横关系得

$$H_r(r,\varphi) = -\frac{jk_z}{k_c} H_{ni} J_n'(k_c r)\cos(n\varphi)$$

$$H_\varphi(r,\varphi) = \frac{jk_z n}{k_c^2 r} H_{ni} J_n(k_c r)\sin(n\varphi)$$

$$E_r(r,\varphi) = \eta_{TE} H_\varphi(r,\varphi)$$

$$E_\varphi(r,\varphi) = -\eta_{TE} H_r(r,\varphi).$$

No.

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圆TM:
$$\begin{cases} \frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial E_z}{\partial r}\right)+\frac{1}{r^2}\frac{\partial^2 E_z}{\partial \varphi^2}+k_c^2E_z=0 \\ E_z\big|_{r=R}=0 \end{cases}$$

$E_z(r,\varphi)=E_{mn}J_n(k_c r)\cos(n\varphi)$　　利用以———

$E_r(r,\varphi)=-\frac{jk_z}{k_c}E_{mn}J_n'(k_c r)\cos(n\varphi)$

$E_\varphi(r,\varphi)=\frac{jk_z n}{k_c r}E_{mn}J_n(k_c r)\sin(n\varphi)$

$H_r(r,\varphi)=-\frac{1}{\eta_{TM}}E_\varphi(r,\varphi)$

$H_\varphi(r,\varphi)=\frac{1}{\eta_{TM}}E_r(r,\varphi)$.

同轴线TEM:　$\nabla_T\cdot\vec{E}(x,y)=0 \Rightarrow \nabla_T^2\psi=\frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial\psi}{\partial r}\right)+\frac{1}{r^2}\frac{\partial^2\psi}{\partial\varphi^2}=0.$

横向无源无旋　$\Rightarrow \frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial\psi}{\partial r}\right)=0.$

$\psi=\frac{V}{\ln(a/b)}\ln r.$

$\therefore \vec{E}_T(x,y)=-\frac{\partial\psi}{\partial r}\hat{r}=-\frac{V}{r\ln(a/b)}\hat{r}$

$\vec{H}_T(x,y)=\frac{1}{\eta_{TEM}}\hat{z}\times\vec{E}_T(x,y)$　　$\because \eta_{TEM}=\frac{\omega\mu}{k}=\sqrt{\frac{\mu}{\varepsilon}}$

$=377(\Omega)$

$=120\pi(\Omega)$

真空

同轴线TE:
$$\begin{cases} \frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial H_z}{\partial r}\right)+\frac{1}{r^2}\frac{\partial^2 H_z}{\partial\varphi^2}+k_c^2H_z=0 \\ \frac{\partial H_z}{\partial r}\big|_{r=a}=\frac{\partial H_z}{\partial r}\big|_{r=b}=0. \end{cases}$$

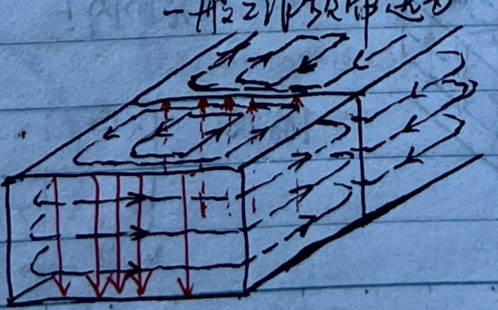
同轴线TM:
$$\begin{cases} \frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial E_z}{\partial r}\right)+\frac{1}{r^2}\frac{\partial^2 E_z}{\partial\varphi^2}+k_c^2E_z=0. \\ E_z\big|_{r=a}=E_z\big|_{r=b}=0. \end{cases}$$

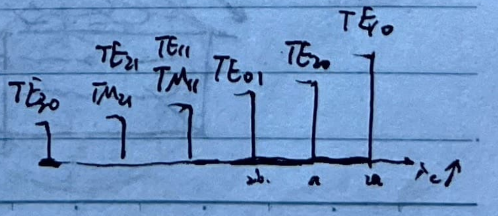
2. 上述3种传输线和微带线中各自传输哪一种主模(基模)? 其主模的电磁力线分布如何? 怎样才能确保主模的单模传输?

矩形:　$TE_{10}^{\square}$　　$\because \lambda_c=2/\sqrt{\left(\frac{m}{a}\right)^2+\left(\frac{n}{b}\right)^2}$, $m=1, n=0 \sim m=2, n=0.$

$\therefore$ 主模TE_{10}的单模工作区为 $a<\lambda_c<2a$, $\frac{\lambda_c}{2}<a<\lambda_c$

一般工作频带选为 $1.05a\le\lambda_c\le 0.8\times 2a$.



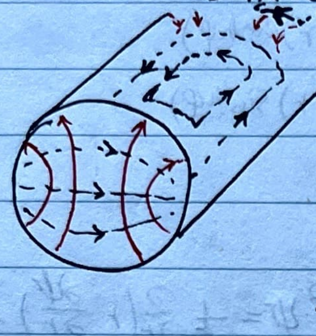


圆： TE_{11}° ∵ $\lambda_c = \frac{2\pi a}{v_{ni}}$, $n=1, i=1 \rightarrow TE_{11}$

$\lambda_c = \frac{2\pi a}{u_{ni}}$, $n=0, i=1 \rightarrow TM_{10}$

∴ 主模 TE_{11} 的单模工作区为 $2.61a < \lambda_c < 3.41a$

$\frac{\lambda_c}{3.41} < a < \frac{\lambda_c}{2.61}$ $a \approx \frac{\lambda}{3}$

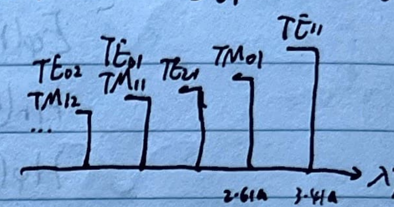


有极化

简并！

(n=0时没有)

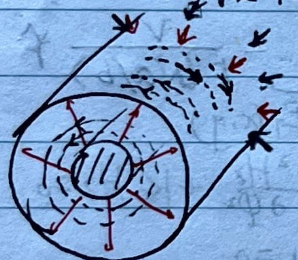
无模式简并！

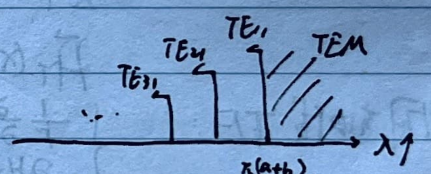


同轴线 $TEM^{\odot}$ ∵ $TE_{11}^{\odot}$: $\lambda_c = \pi(a+b)$ $Z_c = 60\sqrt{\frac{\mu_r}{\varepsilon_r}}\ln\left(\frac{D}{d}\right)$

∴ 主模 TEM 的单模工作区为 $\lambda_c \geq \pi(a+b) \Rightarrow \lambda_c \geq 1.1\pi(a+b)$

$\frac{\pi}{2}(D+d)$

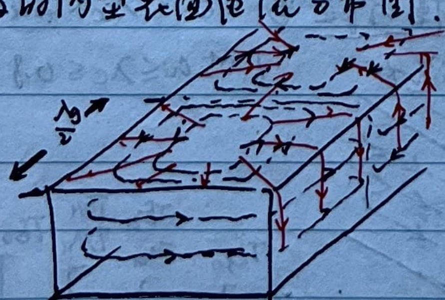




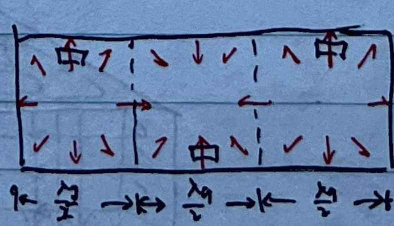
微带线 准TEM. $\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_e}}$

3. $TE_{10}^{\square}$ 波的内壁表面电流分布图，哪些槽缝会切割这些表面电流而引起辐射？

允向内！



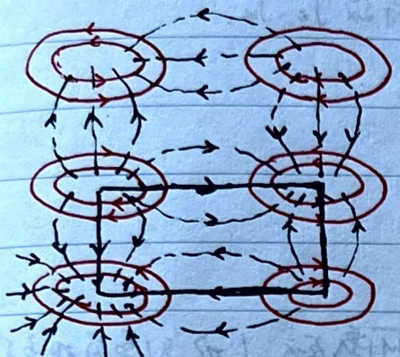
俯视：

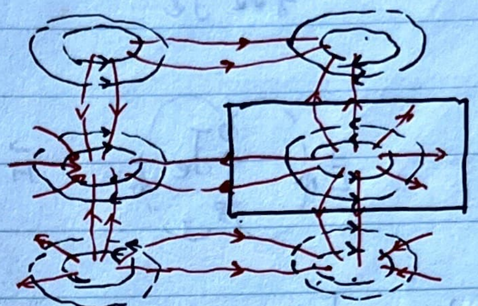


{ TE: E转圈成贴壁
{ TM: H转圈.

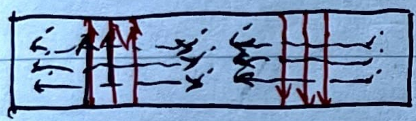
4. 矩形波导和圆波导中高次模的画法.

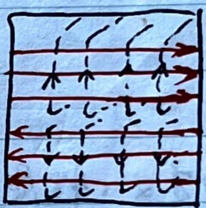
矩形: TE_{mn} TM_{mn}





圆: TE_{m0}. (场分量在y方向均匀分布) TE_{0n}. (场分量在x方向均匀分布).

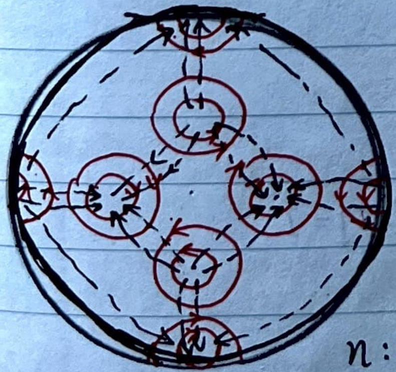




m: 宽边场分量变化的"半周期数"

n: 窄边场分量变化的"半周期数".

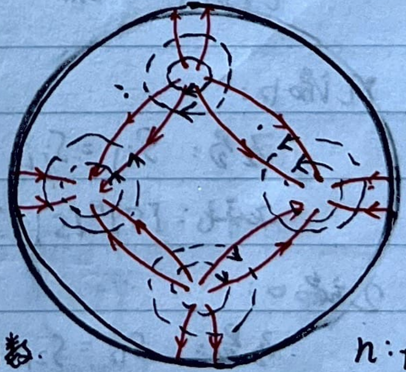
圆:



TE_{ni}

n: 场分量在圆周上变化的周期数.

i: H_z分量在$0<r<a$上的极值数 (不包括0处的极值点).

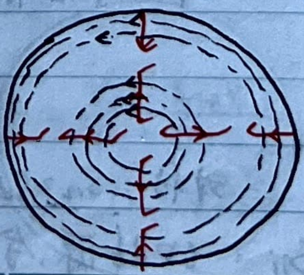


TM_{ni}

n: 场分量在圆周上变化的周期数

i: E_z分量在$0<r<a$上的零点数 (不包括0处的零点).

TE_{0i} (场分量在圆周均匀分布)

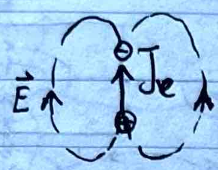


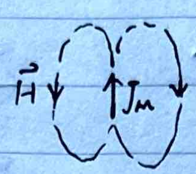
TM_{0i} (场分量在圆周均匀分布).

5. 波导的激励方法与准则.

激励方法： 电场激励、磁场激励、电磁激励

等效 $\vec{J}_e$ 　　等效 $\vec{J}_m$ 　　等效 $\vec{J}_e$、$\vec{J}_m$.





激励准则：（条件） $\int_V (\vec{E}\cdot\vec{J}_e^* + \vec{H}^*\cdot\vec{J}_m)\,dV \begin{cases} \neq 0 & \text{可激励（场与源有相互作用）} \\ = 0 & \text{不可激励（场与源无相互作用）.} \end{cases}$

要注意极化条件和对称条件！

三. 微波网络.

1. 散射参量的定义、性质、物理意义？

定义： $S_{ij} = \frac{b_i}{a_j}\Big|_{a_{k\neq j}=0}$，据左式，$S_{ii}=\Gamma_i$

$$\begin{bmatrix} b_1 \\ \vdots \\ b_n \end{bmatrix} = \begin{bmatrix} S_{11} & \cdots & S_{1n} \\ \vdots & \ddots & \vdots \\ S_{n1} & \cdots & S_{nn} \end{bmatrix} \begin{bmatrix} a_1 \\ \vdots \\ a_n \end{bmatrix} \qquad \begin{cases} a = I_+\sqrt{Z_c} = V_+/\sqrt{Z_c} \\ b = -I_-\sqrt{Z_c} = V_-/\sqrt{Z_c} \end{cases}$$

性质： n端口：

互易：$S_{ij}=S_{ji}$，$[S]=[S]^T$.

无耗：$[S]^H[S]=I$，∴互易无耗 $[S]^*\cdot[S]=I$

2端口：

互易：$S_{12}=S_{21}$ （三未知）

列和为1

无耗：$|S_{12}|=|S_{21}|$，$|S_{11}|=|S_{22}|$，$|S_{11}|^2+|S_{21}|^2=1$，$|S_{12}|^2+|S_{22}|^2=1$（[illegible]未知）.

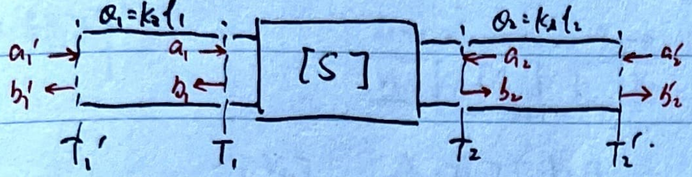
（值：四未知
模：一未知）

对称：$S_{12}=S_{21}$，$S_{11}=S_{22}$（二未知）

物理意义：S_{ii}：i口接源，余口接匹配负载，i口的反射系数.

S_{ij}：j口接源，余口接匹配负载，j至i口的电压传输系数

2. 参考面移动后的 [S]



$$S_{11}' = \frac{b_1 e^{-j\theta_1}}{a_1 e^{j\theta_1}}\Big|_{a_2=0} = S_{11} e^{-j2\theta_1}$$

$$S_{22}' = \frac{b_2 e^{-j\theta_2}}{a_2 e^{j\theta_2}}\Big|_{a_1=0} = S_{22} e^{-j2\theta_2}.$$

$$S_{21}' = \frac{b_2 e^{-j\theta_2}}{a_1 e^{j\theta_1}}\Big|_{a_2=0} = S_{21} e^{-j(\theta_1+\theta_2)}$$

$$S_{12}' = \frac{b_1 e^{-j\theta_1}}{a_2 e^{j\theta_2}}\Big|_{a_1=0} = S_{12} e^{-j(\theta_1+\theta_2)}$$

$$\therefore\ [S'] = \begin{bmatrix} S_{11} e^{-j2\theta_1} & S_{12} e^{-j(\theta_1+\theta_2)} \\ S_{21} e^{-j(\theta_1+\theta_2)} & S_{22} e^{-j2\theta_2} \end{bmatrix} = \begin{bmatrix} e^{-j\theta_1} & 0 \\ 0 & e^{-j\theta_2} \end{bmatrix} \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} e^{-j\theta_1} & 0 \\ 0 & e^{-j\theta_2} \end{bmatrix}$$

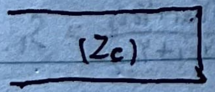
n端口：$[S'] = [P][S][P]$，$[P] = \mathrm{diag}\{e^{-j\theta_i}\}$. $i = 1, \dots, n$.

3. 对称二端口网络 [S] 的求解方法. ∵ 开路 → S_1；短路 → S_2.

P187 例4.5.

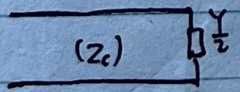
$S_{22} = S_{11} = \frac{S_1+S_2}{2}$，$S_{12} = S_{12} = \frac{S_1-S_2}{2}$.

短：



$S_2 = -1$

开：

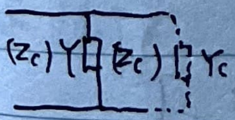


$$S_1 = \frac{-Y/2 \cdot Z_c + 1}{+Y/2 \cdot Z_c + 1} = \frac{2Y_c - Y}{2Y_c + Y}.$$

$$\therefore\ S_{11} = S_{22} = \frac{S_1+S_2}{2} = \frac{1}{2}\cdot\frac{-2Y}{2Y_c+Y} = -\frac{Y}{2Y_c+Y}.$$

$$S_{12} = S_{21} = \frac{S_1-S_2}{2} = \frac{1}{2}\cdot\frac{4Y_c}{2Y_c+Y} = \frac{2Y_c}{2Y_c+Y}$$

按物理意义求：



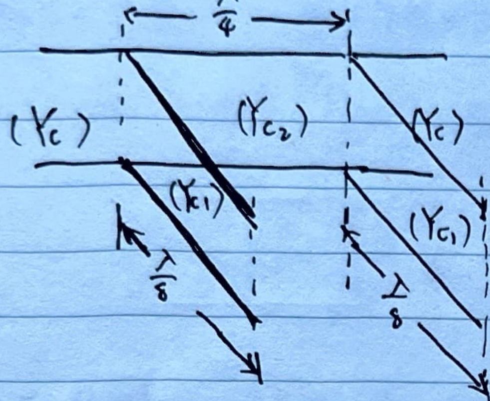
$$S_{11} = \frac{b_1}{a_1}\Big|_{a_2=0} = \Gamma_1 = \frac{1-\bar{y}_1}{1+\bar{y}_1} = \frac{Y_c - (Y+Y_c)}{Y_c + (Y+Y_c)} = -\frac{Y}{Y+2Y_c}.$$

$$S_{21} = \frac{b_2}{a_1}\Big|_{a_2=0} = \frac{a_1 + a_1\Gamma_1}{a_1} = 1 + \Gamma_1 = \frac{2Y_c}{Y+2Y_c}$$

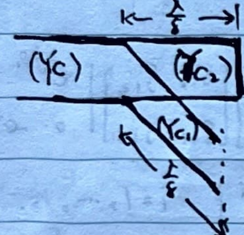
No.

Date

课上例题:



短:



$\Gamma_{\phi 2} = -e^{-jk_2\cdot 2\cdot\frac{\lambda}{8}} = -(-j) = j$, $\bar{Z}_{\phi 2} = \frac{1+\Gamma_{\phi 1}}{1-\Gamma_{\phi 1}} = \frac{1+j}{1-j}$

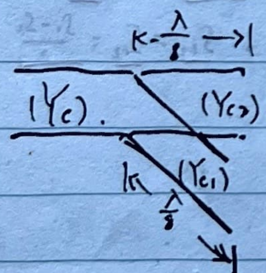
$\Gamma_{\phi 1} = e^{-jk_2\cdot 2\cdot\frac{\lambda}{8}} = -j$, $\bar{Z}_{\phi 1} = \frac{1+\Gamma_{\phi 2}}{1-\Gamma_{\phi 2}} = \frac{1-j}{1+j}$

$\therefore Z_{\phi 2} = \frac{1}{Y_{c2}}\cdot\frac{1+j}{1-j}$, $Z_{\phi 1} = \frac{1}{Y_{c1}}\frac{1-j}{1+j}$

$\therefore Y_\phi = Y_{c2}\frac{1-j}{1+j} + Y_{c1}\frac{1+j}{1-j}$ $\therefore \bar{Y}_\phi = \frac{Y_{c2}}{Y_c}\frac{1-j}{1+j} + \frac{Y_{c1}}{Y_c}\frac{1+j}{1-j}$

$\therefore \Gamma_{左} = \frac{1-\bar{Y}_\phi}{1+\bar{Y}_\phi} = \frac{Y_c-(-jY_{c2}+jY_{c1})}{Y_c+(-jY_{c2}+jY_{c1})} = \frac{Y_c - j(Y_{c1}-Y_{c2})}{Y_c+j(Y_{c1}-Y_{c2})} = S_2$

开:



$\Gamma_{\phi 2} = e^{-jk_2\cdot 2\cdot\frac{\lambda}{8}} = -j$, $\bar{Z}_{\phi 2} = \frac{1+j}{1-j} = j$, $Y_{\phi 2} = +jY_{c2}$.

$\Gamma_{\phi 1} = -j$, $\bar{Z}_{\phi 1} = -j$, $Y_{\phi 1} = +jY_{c1}$

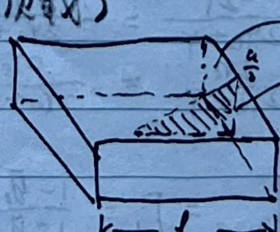
$\therefore Y_\phi = +j(Y_{c1}+Y_{c2})$

$\therefore \Gamma_{左} = \frac{Y_c - Y_\phi}{Y_c + Y_\phi} = \frac{Y_c - j(Y_{c1}+Y_{c2})}{Y_c + j(Y_{c1}+Y_{c2})} = S_1$.

$\therefore S_{11} = S_{22} = \frac{S_1+S_2}{2}$, $S_{12} = S_{21} = \frac{S_1-S_2}{2}$

补充题: (单端口、TE_{10}匹配负载)

$\rightsquigarrow TE_{10}/TE_{20}$.



短路面(对TE_{20})

TE_{10}匹配负载: $S_{11}^{TE_{10}} = 0$

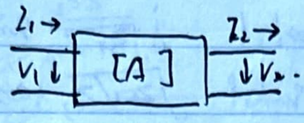
TE_{20}无耗: $S_{11}^{TE_{20}} = -e^{-j2k_2 l}$

混合S阵 $[S] = \begin{bmatrix} S^{TE_{10}} & S^{TE_{10},TE_{20}} \\ S^{TE_{20},TE_{10}} & S^{TE_{20}} \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 0 & -e^{-j2k_2 l} \end{bmatrix}$

↳ TE_{10}输入，接TE_{10}匹配负载时，$\frac{输出的TE_{20}}{输入的TE_{10}}$

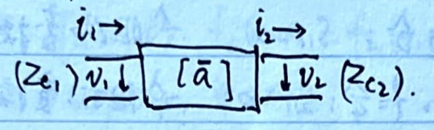
4. 级联矩阵 [A]、[a]   [a]~[S]的换算.

级联矩阵



$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \qquad [A] = \prod_{i=1}^{n} [A_i].$$

归一化级联矩阵（转移矩阵）.



$$\left\{ \begin{aligned} i &= I\cdot\sqrt{Z_c} \\ v &= V/\sqrt{Z_c}. \end{aligned} \right.$$

$$\begin{bmatrix} \bar{a} & \bar{b} \\ \bar{c} & \bar{d} \end{bmatrix} = \begin{bmatrix} A\sqrt{\frac{Z_{c2}}{Z_{c1}}} & B/\sqrt{Z_{c1}Z_{c2}} \\ C\sqrt{Z_{c1}Z_{c2}} & D\sqrt{\frac{Z_{c1}}{Z_{c2}}} \end{bmatrix}.$$

[a]~[S]的换算:

$$S_{11} = \frac{\bar{a}+\bar{b}-\bar{c}-\bar{d}}{\bar{a}+\bar{b}+\bar{c}+\bar{d}} \qquad \bar{a} = \frac{1}{2S_{21}}(1+S_{11}-S_{22}-|S|)$$

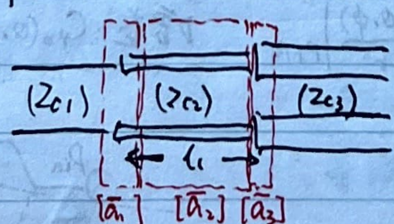
$$S_{12} = \frac{2|\bar{a}|}{\bar{a}+\bar{b}+\bar{c}+\bar{d}} \qquad \bar{b} = \frac{1}{2S_{21}}(1+S_{11}+S_{22}+|S|)$$

$$S_{21} = \frac{2}{\bar{a}+\bar{b}+\bar{c}+\bar{d}} \qquad \bar{c} = \frac{1}{2S_{21}}(1-S_{11}-S_{22}+|S|)$$

$$S_{22} = \frac{-\bar{a}+\bar{b}-\bar{c}+\bar{d}}{\bar{a}+\bar{b}+\bar{c}+\bar{d}} \qquad \bar{d} = \frac{1}{2S_{21}}(1-S_{11}+S_{22}-|S|).$$

例：求$[\bar{a}]$，何时输入端阻抗匹配（$S_{11}=\Gamma_1=0$）？ 同质用$\bar{a}$、异质用$\bar{A}$方便.

需对Z_c归一化　无需对Z_c归一化.



首先求$[\bar{a}_2]$： $\bar{a} = \frac{v_1}{v_2}\Big|_{i_2=0} = \frac{V_2\cos k_2 l}{V_2} = \cos k_2 l$

$\bar{b} = \frac{v_1}{i_2}\Big|_{v_2=0} = \frac{j2V_{0+}\sin k_2 l}{2V_{0+}} = j\sin k_2 l.$

∵对称 ∴ $\bar{a} = \bar{d}$，$\bar{b} = \bar{c}$.

再求$[\bar{a}_1]$：先求$[A_1]$：（非对称）

$\bar{A} = \frac{V_1}{V_2}\Big|_{I_2=0} = 1$，$\bar{B} = \frac{V_1}{I_2}\Big|_{V_2=0} = 0$，$\bar{C} = \frac{I_1}{V_2}\Big|_{I_2=0} = 0$，$\bar{D} = \frac{I_1}{I_2}\Big|_{V_2=0} = 1.$

$$\therefore [\bar{a}_1] = \begin{bmatrix} \sqrt{\frac{Z_{c2}}{Z_{c1}}}\bar{A} & \bar{B}/\sqrt{Z_{c1}Z_{c2}} \\ \sqrt{Z_{c1}Z_{c2}}\bar{C} & \bar{D}\sqrt{\frac{Z_{c1}}{Z_{c2}}} \end{bmatrix} = \begin{bmatrix} \sqrt{\frac{Z_{c2}}{Z_{c1}}} & 0 \\ 0 & \sqrt{\frac{Z_{c1}}{Z_{c2}}} \end{bmatrix}.$$

另外，$[\bar{a}_3]$与$[\bar{a}_1]$类似.

$\therefore [\bar{a}] = [\bar{a}_1]\cdot[\bar{a}_2]\cdot[\bar{a}_3]$

$$= \begin{bmatrix} (Z_{c3}/Z_{c1})\cos k_2 l & j\frac{Z_{c2}}{\sqrt{Z_{c1}Z_{c3}}}\sin k_2 l \\ \sqrt{Z_{c3}/Z_{c1}}\sin k_2 l & \sqrt{Z_{c1}/Z_{c3}}\cdot\cos k_2 l \end{bmatrix}$$

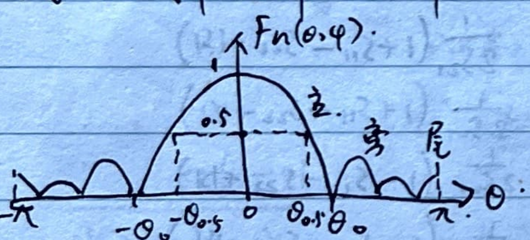
换算出 S_{11}： $S_{11} = \frac{\bar{a}+\bar{b}-\bar{c}-\bar{d}}{\bar{a}+\bar{b}+\bar{c}+\bar{d}} = \frac{1}{k}\left[\left(\sqrt{Z_{c3}/Z_{c1}}-\sqrt{Z_{c1}/Z_{c3}}\right)\cos k_2 l + j\left(\frac{Z_{c2}}{\sqrt{Z_{c1}Z_{c3}}}-\frac{\sqrt{Z_{c1}Z_{c3}}}{Z_{c2}}\right)\sin k_2 l\right]$

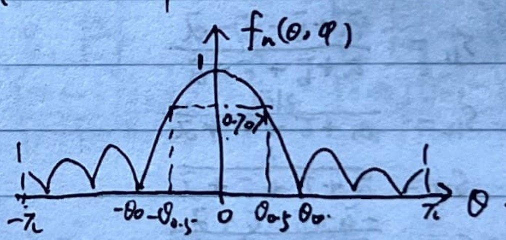
$(k\neq 0)$.

$$\begin{cases} Z_{c1}=Z_{c2}=Z_{c3} : S_{11}=0 & \text{均匀长线} \\ Z_{c1}=Z_{c3},\ l=\frac{\lambda}{2} : S_{11}=0 & \frac{\lambda}{2}\text{阻抗变换器} \\ Z_{c2}=\sqrt{Z_{c1}Z_{c3}},\ l=\frac{\lambda}{4} : S_{11}=0 & \frac{\lambda}{4}\text{阻抗变换器} \end{cases}$$

四、天线

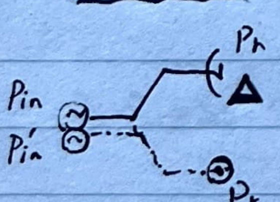
1. 天线方向图中主瓣、旁瓣、尾瓣、半功率角、主瓣零点张角的含义.

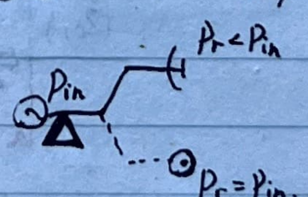




2. 天线方向性系数 D 和增益 G 的定义？它们有何区别？

方向性系数：$D_0(\theta,\varphi) = \frac{S_{r,max}(\theta,\varphi)}{P_r/4\pi r^2}\Big|_{P_r同}$ 增益：$G_0(\theta,\varphi) = \frac{S_{r,max}(\theta,\varphi)}{P_r/4\pi r^2}\Big|_{P_{in}同}$





理想天线无耗：$\boxed{G_0 = D_0}$ 一般实际天线：$\boxed{G_0 < D_0}$

3. 何谓天线的极化、极化匹配、极化正交？

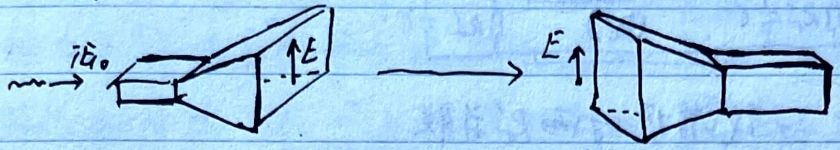
极化（最大增益方向）{
- 线极化：平行（E平行于接收端面）、垂直（垂直于平行极化）
- 圆极化：右旋（右手螺旋）、左旋（左手螺旋）.
- 椭圆极化：等于两反旋向的不等振幅圆极化波的叠加.

No.

Date

极化匹配：收发天线极化一致.

极化正交：极化失配.

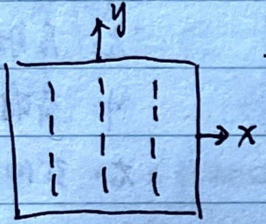


4. 何谓阵列天线方向图相乘原理？平面阵列天线的总方向图的求解思路？

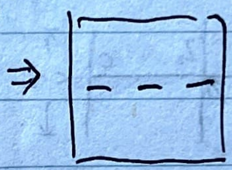
$$f_n(\theta,\varphi) = f_a(\theta,\varphi)\cdot f_e(\theta,\varphi) \approx f_a(\theta,\varphi)$$

阵列因子. 单元因子

平面：



先求 $f_{a,y}(\theta,\varphi)$ ⇒



再求 $f_{a,x}(\theta,\varphi)$

$\therefore f_n(\theta,\varphi) \approx f_{a,y}(\theta,\varphi)\cdot f_{a,x}(\theta,\varphi)$.

5. 全波振子、半波振子、电偶极子各自的E面和H面方向图（归一化场强）？

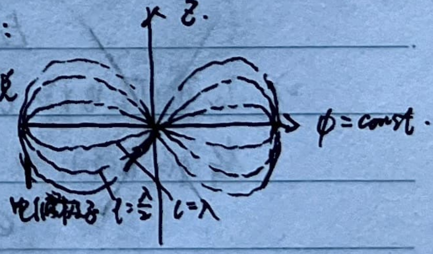
电偶极子：$f_n(\theta)=\sin\theta$

半波振子：$f_n(\theta)=\dfrac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}$.

全波振子：$f_n(\theta)=\dfrac{\cos(\pi\cos\theta)+1}{\sin\theta}$.

• E面（归一化场强）：

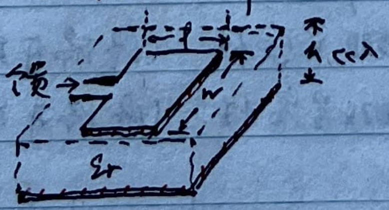
$l<\lambda$ 时，l 越大波瓣越锐



• H面（归一化）：圆.

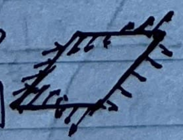
6. 矩形微带天线的结构、分析方法、等效电路.

结构：

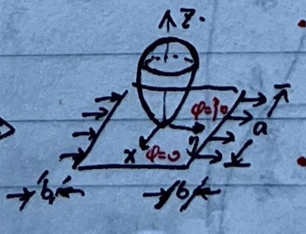


分析方法.

以矩形口面天线为单元的二元线阵列天线.



⇒



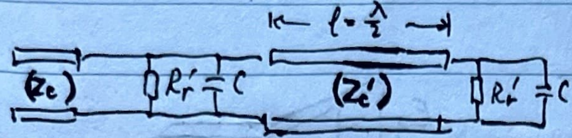
• E面：$\varphi=90^\circ$（有 $\frac{kL}{2}\sin\theta$ 波程差）

$$f_{a,E} = S_a\left(\frac{kb}{2}\sin\theta\right)\cdot 2\cos\left(\frac{kL}{2}\sin\theta\right)$$

• H面：$\varphi=0^\circ$（无波程差，加倍）.

$$f_{a,H} = S_a\left(\frac{ka}{2}\sin\theta\right)2\cos\theta.$$

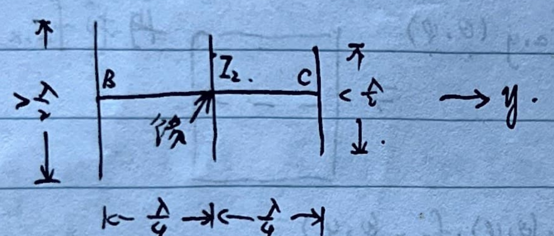
等效电路.



$Z_{in}=R_r/2$. ∵天线谐振时两R_r并联.

{调R_r：改变W
{调λ：改变l.

7. 三单元八木天线的分析方法. $\overset{0}{I_2}\to\overset{0}{E_B}、\overset{0}{E_C}\to\overset{-\pi}{\vec E_B'}、\overset{-\pi}{\vec E_C'}\to\overset{-\pi}{e_B}、\overset{-\pi}{e_C}\xrightarrow[Z_B\quad Z_C]{\frac{\pi}{2}\quad-\frac{\pi}{2}}\overset{\frac{\pi}{2}}{I_1}、\overset{-\frac{\pi}{2}}{I_3}$



左：反射器

中：有源振子

右：引向器

1) I_2在B、C的辐射场（近似远场）

$$E_{B,C}=j\eta\frac{I_2e^{-jkr}}{2\pi r}\cdot\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}=\frac{240I_2}{\lambda}\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}.\quad l=\frac{\lambda}{2}.$$

$E_B、E_C$与I_2同相.

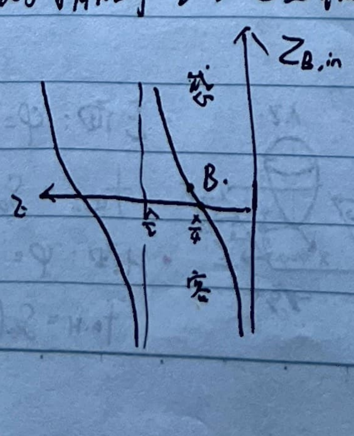
2) I_2感应电场$\vec E_B、\vec E_C$ // 左右两线，在线内的感生电场$\vec E_B'、\vec E_C'$与$\vec E_B、\vec E_C$等幅反相

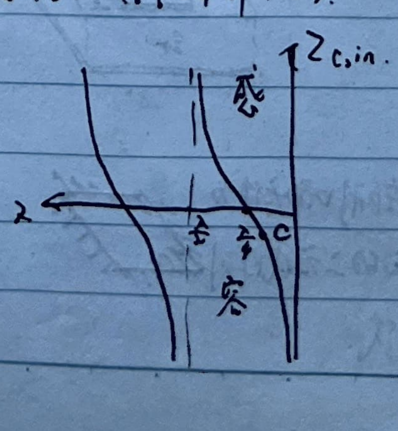
$$e_B=\int_l \vec E_B'\cdot d\vec l=-\int_l \vec E_B\cdot d\vec l$$

$$e_C=\int_l \vec E_C'\cdot d\vec l=-\int_l \vec E_C\cdot d\vec l$$

3). 左振子稍长于$\frac{\lambda}{2}$，B处输入电阻呈感性（终端开路）

右振子稍短于$\frac{\lambda}{2}$，C处输入电阻呈容性.（终端开路）.





4) 由 $I_B = e_B/Z_{B,in}$，$I_C = e_C/Z_{C,in}$，I_B ~~滞后~~ 领先 I_2 90°，I_2 ~~滞后~~ 领先 I_C 90°。

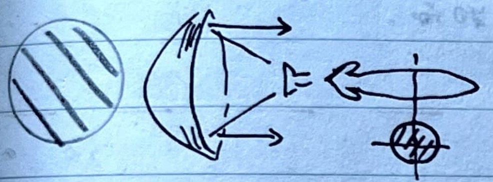
又由近场感应耦合：$I_B \approx I_2 \approx I_C$.

5). 左振子和中振子在 $\hat{y}$ 远场同相叠加，在 $-\hat{y}$ 反相叠加。

右振子和中振子在 $\hat{y}$ 远场同相叠加，在 $-\hat{y}$ 反相叠加。

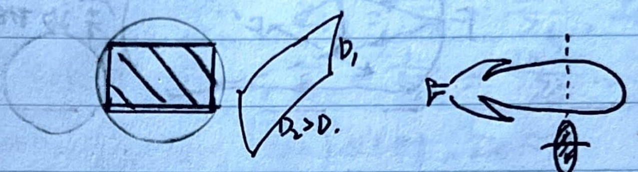
8. 试述几种形式的抛物面天线，它们产生何种波束，作何用途？

a) 旋转抛物面天线（圆形口面） b) 切割抛物面天线（矩形口面）



针状波束

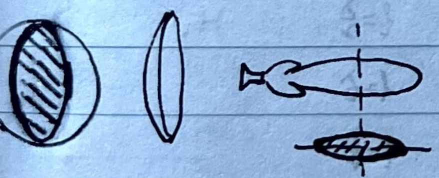
跟踪雷达，卫星地面站



垂直扇形波束

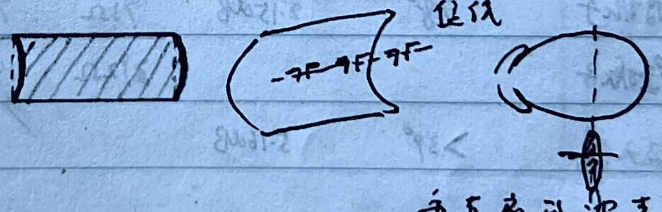
警戒雷达，防航舰

c). 切割抛物面天线（椭圆形口面） d) 抛物柱面天线（矩形口面）



水平扇形波束

测高雷达：通过测俯仰角，b.c合用



垂直扇形波束

海用导航雷达

9. 求抛物面天线远区场的方法与思路？

口面场法：由初级波瓣和抛物面反射特性求焦面上的 $\vec{E_a}$、$\vec{H_a}$（只存于孔径面）

$\boxed{H_a E_a \to J'_{ms} J'_{es} \to A_m A \to E \cdot H}$

由电场源等效原理 & 镜像定理求 $\vec{J'_{es}}$ $\vec{J'_{ms}}$

由 $\vec{J'_{ms}}$ 求 $\vec{A_m}$、$\vec{J'_{es}}$ 求 $\vec{A}$.

由 $\vec{A_m}$、$\vec{A}$ 求 $\vec{E}$、$\vec{H}$.

镜面电流法：由抛物反射面内表面的传导电流求远区场，$\vec{J_s} = 2(\hat{n} \times \vec{H_i})$.

$\boxed{H_i \to J_s \to A \to E \cdot H}$

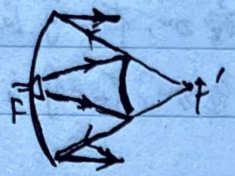
由初级波瓣特性求 $\vec{H_i}$，即得 $\vec{J_s}$

由 $\vec{J_s}$ 求 $\vec{A}$，积分在口面进行

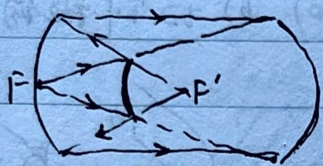
由 $\vec{A}$ 求 $\vec{E}$、$\vec{H}$.

10. 卡塞格伦天线的两种分析方法.

虚馈源法:(等效馈源法)



等效抛物面法: 给定双曲面和抛物面, 只有唯一的等效抛物面相对应.

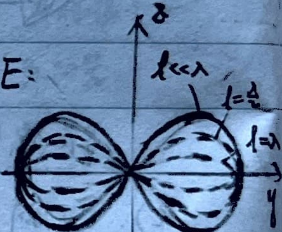


等效抛物面.

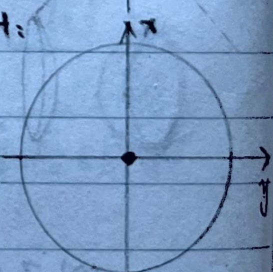
17. 附加

	$2\theta_{0.5}$	D_0	Z_{in}	R_r	l
电流元	90°	1.76dB		$80\pi^2(\frac{l}{\lambda})^2$	$l \ll \lambda$
全波振子	47.8°			200Ω	$l=\lambda$
半波振子	78°	2.15dB	73Ω	73Ω	$l=\frac{\lambda}{2}$
折叠振子			292Ω	146Ω	$l=\frac{\lambda}{2}$
单极天线	>39°	5.16dB			$l=\frac{\lambda}{4}$

E:



H:



模式分类:

1) 横电波 TE波 H波 磁波 $\eta_{TE}=\frac{\omega\mu}{k}$

2) 横磁波 TM波 E波 电波 $\eta_{TM}=\frac{k}{\omega\varepsilon}$

3) 横电磁波 TEM波 $\eta_{TEM}=\frac{\omega\mu}{k}=\frac{k}{\omega\varepsilon}=\sqrt{\frac{\mu}{\varepsilon}}\overset{真空}{=}120\pi=377(\Omega)$

4) 混合波 EH波 HE波: TE波和TM波的线性组合. e.g. 简并.

模式传播条件:

1) 电磁激励方式: 激励与耦合 { 1) 激励工作模式, 抑制非工作模式 2) 阻抗匹配 3) 馈入能量大小可调

2) 工作波长: λ

3) 波导尺寸: λ_c

无源微波电路.

1. 匹配负载

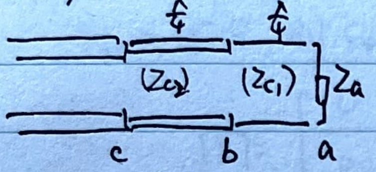
TE₁₀：平行 E_y：尖劈形吸收片.

2. 波导接头

TE₁₀：波导宽壁中线附近有较强 I_z，容易好接触.

3. 短路器

S形扼流活塞：



$$\bar{Z}_a=\frac{Z_a}{Z_{c1}}，\bar{Z}_{b右}=\frac{1}{\bar{Z}_a}=\frac{Z_{c1}}{Z_a}，Z_b=\frac{Z_{c1}^2}{Z_a}$$

$$\bar{Z}_{b右}=\frac{Z_{c1}^2}{Z_aZ_{c2}}，\bar{Z}_{c右}=\frac{1}{\bar{Z}_{b右}}=\frac{Z_aZ_{c2}}{Z_{c1}^2}，Z_c=\frac{Z_aZ_{c2}^2}{Z_{c1}^2}$$

当 $Z_a\approx0$，$\left(\frac{Z_{c2}}{Z_{c1}}\right)^2\ll1$ 时，$Z_c\approx0$.

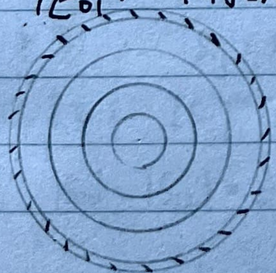
4. 衰减器.

旋转极化式衰减器：$TE_{10}^{\square}\to TE_{11}^{\circ}\cdot\cos\theta\to TE_{10}^{\square}\cdot\cos^2\theta$

$$[S]=\begin{bmatrix}0&\cos^2\theta\\\cos^2\theta&0\end{bmatrix}$$

5. 模式抑制器

TE_{01}° 的抑制器



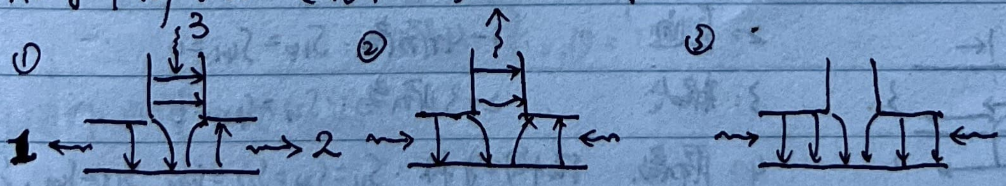
对 TE_{10}°：

$$[S]=\begin{bmatrix}-1&0\\0&-1\end{bmatrix}$$

对通过模式：

$$[S]=\begin{bmatrix}e^{-jkl}&0\\0&e^{-jkl}\end{bmatrix}$$

6. 波导T形分支：（无耗互易三端口网络）



结构对称：$S_{11}=S_{22}$；互易：$S_{ij}=S_{ji}$；结构反对称：$S_{23}=-S_{13}$；无耗：$[S]^{+}[S]=I$，得

$$[S]=\frac{1}{2}\begin{bmatrix}1&1&\sqrt{2}\\1&1&-\sqrt{2}\\\sqrt{2}&-\sqrt{2}&0\end{bmatrix}$$

无耗互易三端口性质：

① 不可能三端口同时匹配

② 不可能两端口同时匹配，否则退化为二端口.

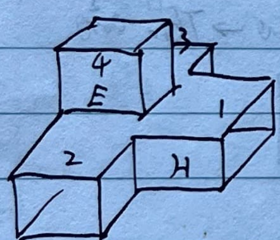
7. 3dB微带功分器（有耗三端口网络）.

$$[S]=\begin{bmatrix} 0 & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & 0 & 0 \\ \frac{1}{\sqrt{2}} & 0 & 0 \end{bmatrix}$$

$1\to2、3$：功分器

$2、3\to1$：合成器.

8. 魔T分支.（无耗互易四端口网络）.



结构对称性：$S_{21}=S_{31}$，$S_{22}=S_{33}$

$S_{41}=S_{14}=0$，$S_{24}=-S_{34}$

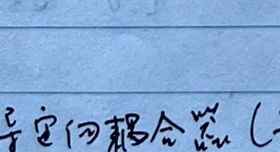
互易性：$S_{ij}=S_{ji}$

1、4口匹配：$S_{11}=S_{44}=0$

无耗：$[S]^H[S]=I$.

得 $[S]=\frac{1}{\sqrt{2}}\begin{bmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & 1 \\ 1 & 0 & 0 & -1 \\ 0 & 1 & -1 & 0 \end{bmatrix}$

9. 波导定向耦合器（无耗互易四端口网络）.



2：直通　　1-4隔离：$S_{14}=S_{41}=0$

3：耦合　　2-3隔离：$S_{23}=S_{32}=0$.

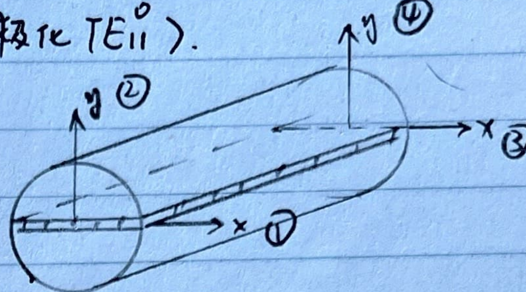
4：隔离.　结构对称：$S_{11}=S_{22}=S_{33}=S_{44}$，$S_{13}=S_{42}$，$S_{12}=S_{4}$

互易、无耗 $\Rightarrow$

$[S]=\begin{bmatrix} 0 & S_{12} & S_{13} & 0 \\ S_{12} & 0 & 0 & S_{13} \\ S_{13} & 0 & 0 & S_{12} \\ 0 & S_{13} & S_{12} & 0 \end{bmatrix}$ · S_{12}和S_{13}相差$90°$.

10. 圆极化器：（无耗互易四端口网络）.

$\begin{cases}\text{几何特性上：二端口}\\ \text{电特性上：四端口（水平极化 } TE_{11}^{o}\text{，垂直极化 } TE_{11}^{o}\text{）.}\end{cases}$



各端口匹配：$S_{ii}=0$

正交极化：$S_{12}=S_{14}=S_{23}=S_{34}=0$.

互易：$S_{ij}=S_{ji}$

垂直极化领先水平极化 $\frac{\pi}{2}$：$\cancel{S_{31}}\ S_{42}=je^{-j\varphi}$

设 $S_{31}=e^{-j\varphi}$，则 $S_{42}=je^{-j\varphi}$

$$[S]=\begin{bmatrix}0 & 0 & e^{-j\varphi} & 0\\ 0 & 0 & 0 & je^{-j\varphi}\\ e^{-j\varphi} & 0 & 0 & 0\\ 0 & je^{-j\varphi} & 0 & 0\end{bmatrix}$$

11. 旋转对称五端口结.（无耗互易五端口网络）

各端口匹配：$S_{ii}=0$

旋转对称性：$S_{12}=S_{23}=S_{34}=S_{45}=S_{51}$ ；$S_{13}=S_{14}=S_{24}=S_{25}=S_{35}$

互易：$S_{ij}=S_{ji}$

无耗：$[S]^H[S]=I_5$

$\Rightarrow\ |S_{12}|=|S_{13}|=\frac{1}{2}$　设 $S_{12}=\frac{1}{2}$，$S_{13}=\frac{1}{2}e^{j\theta}$，则由

$$S_{13}^{*}S_{12}+S_{12}^{*}S_{13}+|S_{12}|^2=0$$

得

$$\frac{1}{4}e^{j\theta}+\frac{1}{4}e^{-j\theta}+\frac{1}{4}=0$$

∴ $\cos\theta=-\frac{1}{2}$，$\theta=120°$.

可用作四等分功分器.

12. 环行器.（旋转对称无耗非互易三端口网络）.

$[S]=\begin{bmatrix}0 & 0 & 1\\ 1 & 0 & 0\\ 0 & 1 & 0\end{bmatrix}$ 或 $\begin{bmatrix}0 & 1 & 0\\ 0 & 0 & 1\\ 1 & 0 & 0\end{bmatrix}$：功率单向循环传输，反向隔离.