Lecture – 6 Date: 21.08.2014

- Lossy Transmission Line
- Introduction to Smith Chart: The complex Γ– plane
- Transformations on the complex Γ– plane
- Mapping Z to Γ
- Smith Chart – Construction
- Smith Chart – Geography

Lossy Transmission Lines

- Recall that we have been **approximating** low-loss transmission lines as lossless (R =G = 0):

$$\alpha = 0 \qquad \beta = \omega\sqrt{LC}$$

- But, **long** low-loss lines require a **better** approximation:

$$\alpha = \frac{1}{2}\left(\frac{R}{Z_0} + GZ_0\right) \qquad \beta = \omega\sqrt{LC}$$

- Now, if we have **really long** transmission lines (e.g., long distance communications), we can apply **no** approximations at all:

$$\alpha = \mathrm{Re}\{\gamma\} \qquad \beta = \mathrm{Im}\{\gamma\}$$

For these **very** long transmission lines, we find that $\beta = Im\{\gamma\}$ is a **function** of signal **frequency** ω . This results in an extremely serious problem—signal **dispersion**.

Lossy Transmission Lines (contd.)

- Recall that the **phase velocity** $\boldsymbol{v_p}$ (i.e., propagation velocity) of a wave in a transmission line is:

$$v_p = \frac{\omega}{\beta}$$

$$\beta = \mathrm{Im}\{\gamma\} = \mathrm{Im}\left\{\sqrt{(R + j\omega L)(G + j\omega C)}\right\}$$

Thus, for a lossy line, the phase velocity $\boldsymbol{v_p}$ is a function of frequency ω (i.e., $\boldsymbol{v_p(\omega)}$)—this is **bad**!

- Any signal that carries significant **information** must has some non-zero **bandwidth**. In other words, the signal energy (as well as the information it carries) is **spread** across many frequencies.
- If the different frequencies that comprise a signal travel at different velocities, that signal will arrive at the end of a transmission line **distorted**. We call this phenomenon signal **dispersion.**
- Recall for **lossless** lines, however, the phase velocity is **independent** of frequency—**no** dispersion will occur!

Lossy Transmission Lines (contd.)

- For lossless line: $$v_p = \frac{1}{\sqrt{LC}}$$

however, a perfectly lossless line is impossible, but we find phase velocity is **approximately** constant if the line is low-loss.

Therefore, dispersion distortion on low-loss lines is **most often** not a problem.

Q: You say "**most** often" not a problem—that phrase seems to imply that dispersion sometimes **is** a problem!

Lossy Transmission Lines (contd.)

A: Even for low-loss transmission lines, dispersion can be a problem **if** the lines are **very** long—just a small difference in phase velocity can result in significant differences in propagation delay **if** the line is very long!

- Modern examples of long transmission lines include phone lines and cable TV. However, the **original** long transmission line problem occurred with the **telegraph**, a device invented and implemented in the 19th century.
- Early telegraph "engineers" discovered that if they made their telegraph lines **too long**, the dots and dashes characterizing Morse code turned into a muddled, indecipherable **mess**. Although they did not realize it, they had fallen victim to the heinous effects of **dispersion**!
- Thus, to send messages over long distances, they were forced to implement a series of intermediate "**repeater**" stations, wherein a human operator received and then **retransmitted** a message on to the next station. This **really** slowed things down!

Lossy Transmission Lines (contd.)

Q: Is there any way to **prevent** dispersion from occurring?

A: You bet! **Oliver Heaviside** figured out how in the **19th** Century!

- Heaviside found that a transmission line would be distortionless (i.e., no dispersion) **if** the line parameters exhibited the following **ratio**:

$$\frac{R}{L} = \frac{G}{C}$$

- Let's see **why** this works. Note the complex propagation constant γ can be expressed as:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} = \sqrt{LC(R/L + j\omega)(G/C + j\omega)}$$

Lossy Transmission Lines (contd.)

- Then **IF**: $$\frac{R}{L} = \frac{G}{C}$$
- we find:

$$\gamma = \sqrt{LC\left(R/L + j\omega\right)\left(R/L + j\omega\right)} = \left(R/L + j\omega\right)\sqrt{LC} = R\sqrt{\frac{C}{L}} + j\omega\sqrt{LC}$$

- Thus: $$\alpha = \text{Re}\{\gamma\} = R\sqrt{\frac{C}{L}}$$ $$\beta = \text{Im}\{\gamma\} = \omega\sqrt{LC}$$
- The propagation **velocity** of the wave is thus: $$v_p = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}}$$

The propagation velocity is **independent** of frequency! This lossy transmission line is **not** dispersive!

Lossy Transmission Lines (contd.)

Q: Right. All the transmission lines I use have the property that $R/_L > G/_C$. I've **never** found a transmission line with this **ideal** property $R/_L = G/_C$!

A: It is true that typically $R/_L > G/_C$. But, we can reduce the ratio $R/_L$(until it is equal to $G/_C$) by adding series **inductors** periodically along the transmission line.

This was **Heaviside's** solution—and it worked! **Long** distance transmission lines were made possible.

Q: Why don't we increase G instead?

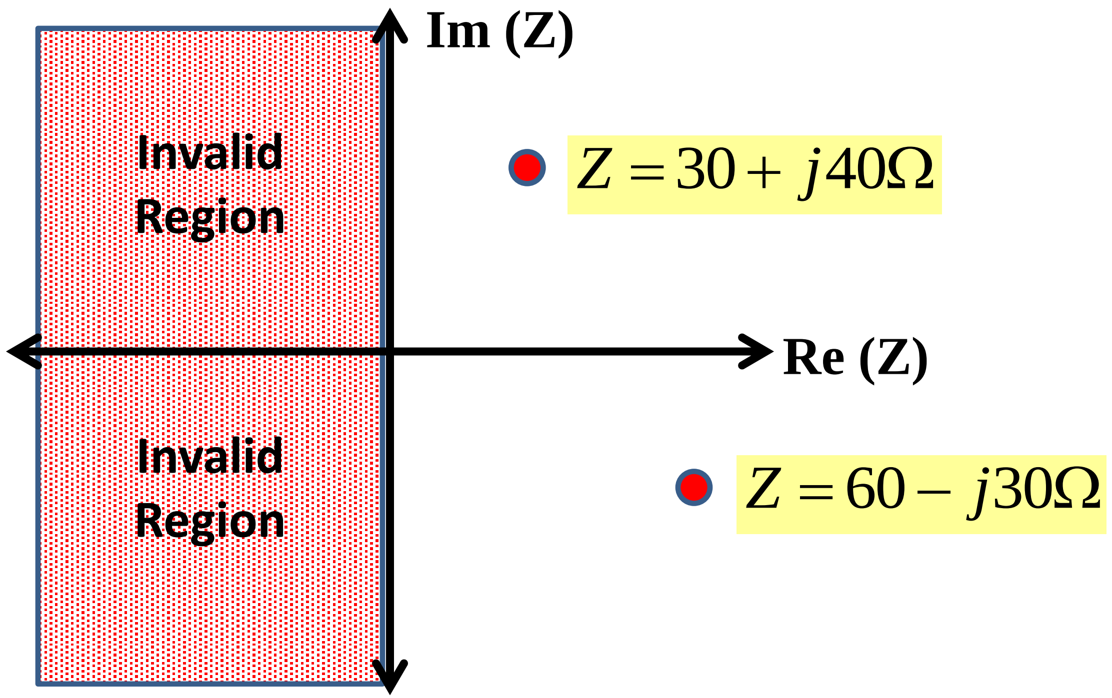
A:

Smith Chart

- Smith chart – what?
- The Smith chart is a very convenient graphical tool for analyzing TLs studying their behavior.
- It is mapping of impedance in standard complex plane into a suitable complex reflection coefficient plane.
- It provides graphical display of reflection coefficients.
- The impedances can be directly determined from the graphical display (ie, from Smith chart)
- Furthermore, Smith charts facilitate the analysis and design of complicated circuit configurations.

The Complex Γ- Plane

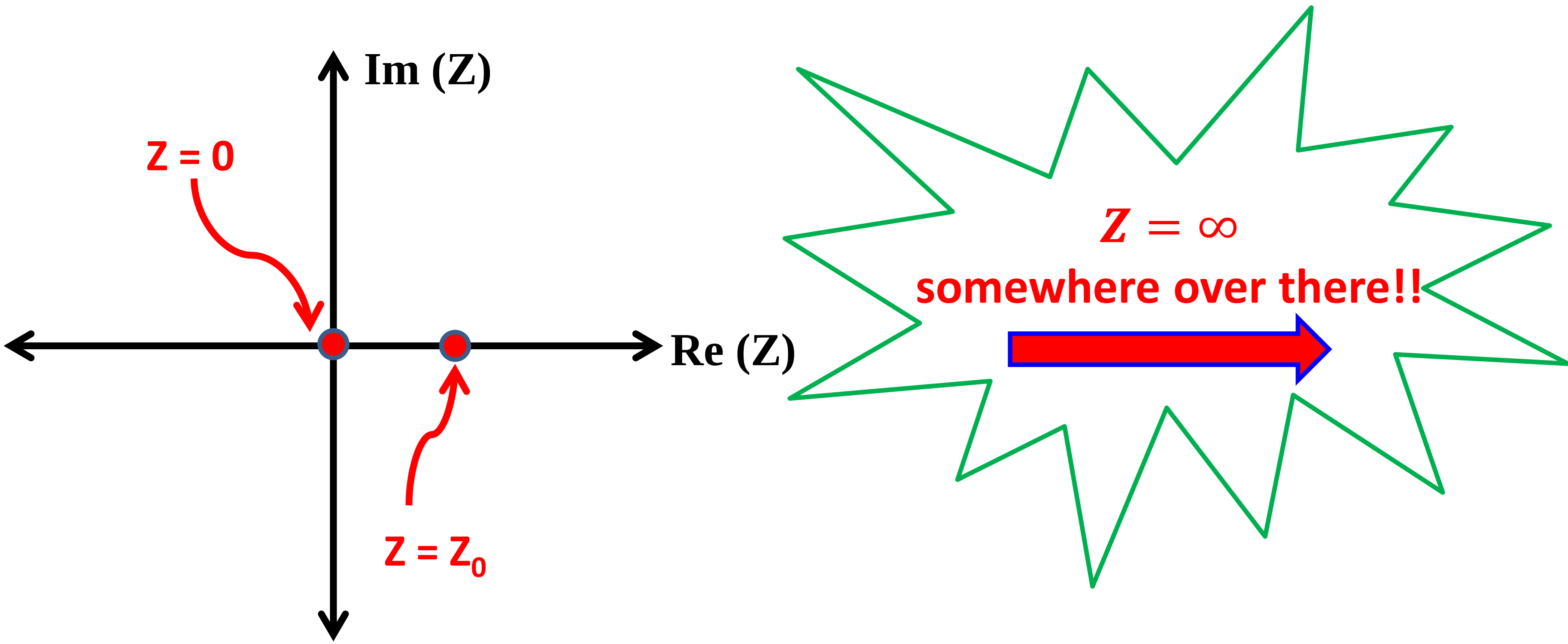
- Let us first display the impedance Z on complex Z-plane



- Note that each dimension is defined by a single real line: the horizontal line (axis) indicates the real component of Z, and the vertical line (axis) indicates the imaginary component of Z → Intersection of these lines indicate the complex impedance

The Complex Γ- Plane (contd.)

- How do we plot an **open circuit** (i.e, $Z = \infty$), **short circuit** (i.e, $Z = 0$), and **matching condition** (i.e, $Z = Z_0 = 50\Omega$) on the complex Z-plane



It is apparent that **complex** $Z - plane$ is not very useful

The Complex Γ-Plane (contd.)

- The **limitations** of complex Z-plane can be **overcome** by complex Γ-plane
- We know **Z ↔ Γ** (i.e, if you know **one**, you know the **other**).
- We can therefore define a complex Γ-plane in the same manner that we defined a complex Z-plane.
- Let us revisit the reflection coefficient in complex form:

$$\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0} = \Gamma_{0r} + \Gamma_{0i} = |\Gamma_0| e^{j\theta_0}$$

Real part of Γ_0: Γ_{0r}; Imaginary part of Γ_0: Γ_{0i}

Where,

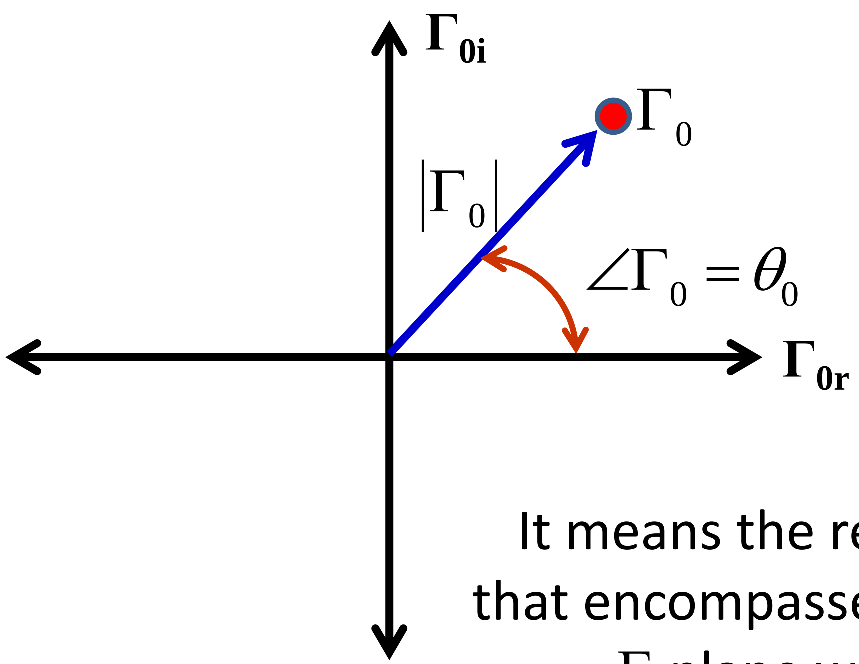
$$\theta_0 = \tan^{-1}\left(\frac{\Gamma_{0i}}{\Gamma_{0r}}\right)$$

- In the special terminated conditions of pure short-circuit and pure open-circuit conditions the corresponding Γ_0 are -1 and +1 located on the real axis in the complex Γ-plane.

The Complex Γ-Plane (contd.)

$$\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0} = \Gamma_{0r} + \Gamma_{0i} = |\Gamma_0| e^{j\theta_0}$$

Representation of reflection coefficient in polar form



Observations:

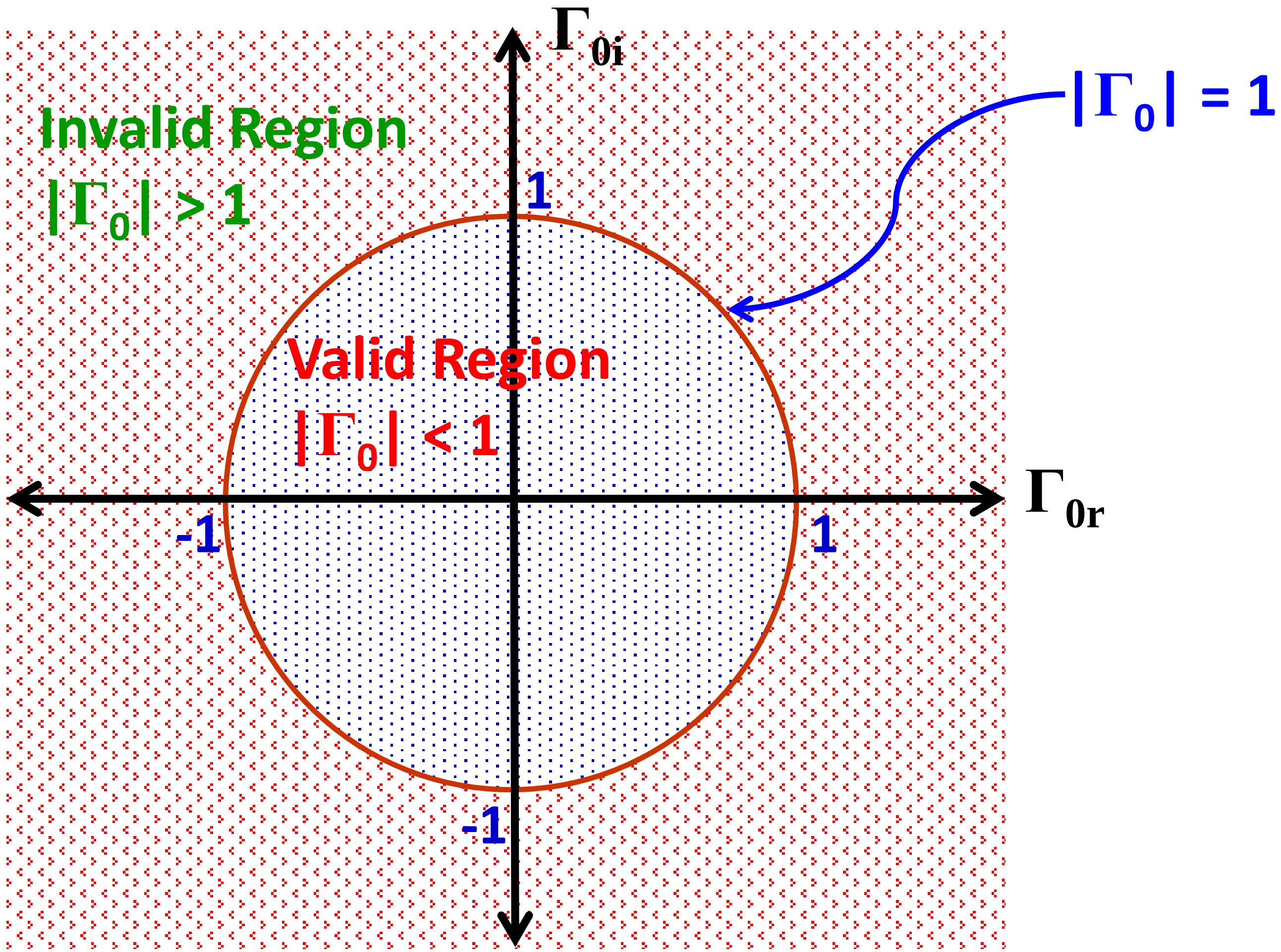
- A radial line is formed by the locus of all points whose phase is θ_0
- A circle is formed by the locus of all points whose magnitude is $|\Gamma_0|$

It means the reflection coefficient has a valid region that encompasses all the four quadrants in the complex Γ-plane within the -1 to +1 bounded region

In complex Z-plane the valid region was unbounded on the right half of the plane → as a result many important impedances could not be plotted

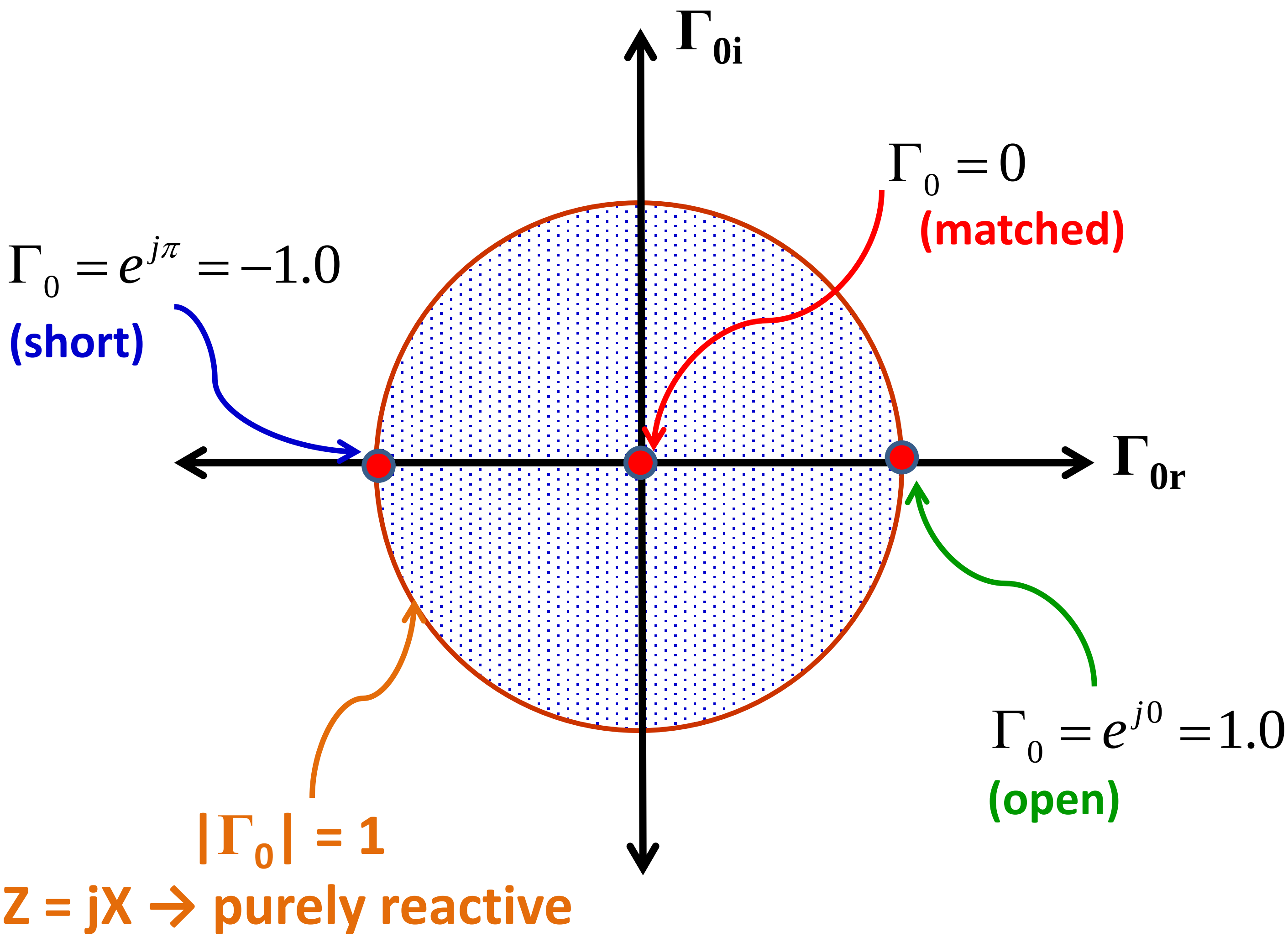
The Complex Γ-Plane (contd.)

- **Validity Region**



The Complex Γ-Plane (contd.)

- We can plot all the valid impedances (i.e R > 0) within this bounded region.



Example – 1

- A TL with a characteristic impedance of Z_0 = 50Ω is terminated into following load impedances:
  - (a) Z_L = 0 (Short Circuit)
  - (b) $Z_L \rightarrow \infty$ (Open Circuit)
  - (c) Z_L = 50Ω
  - (d) Z_L = (16.67 – j16.67)Ω
  - (e) Z_L = (50 + j50)Ω

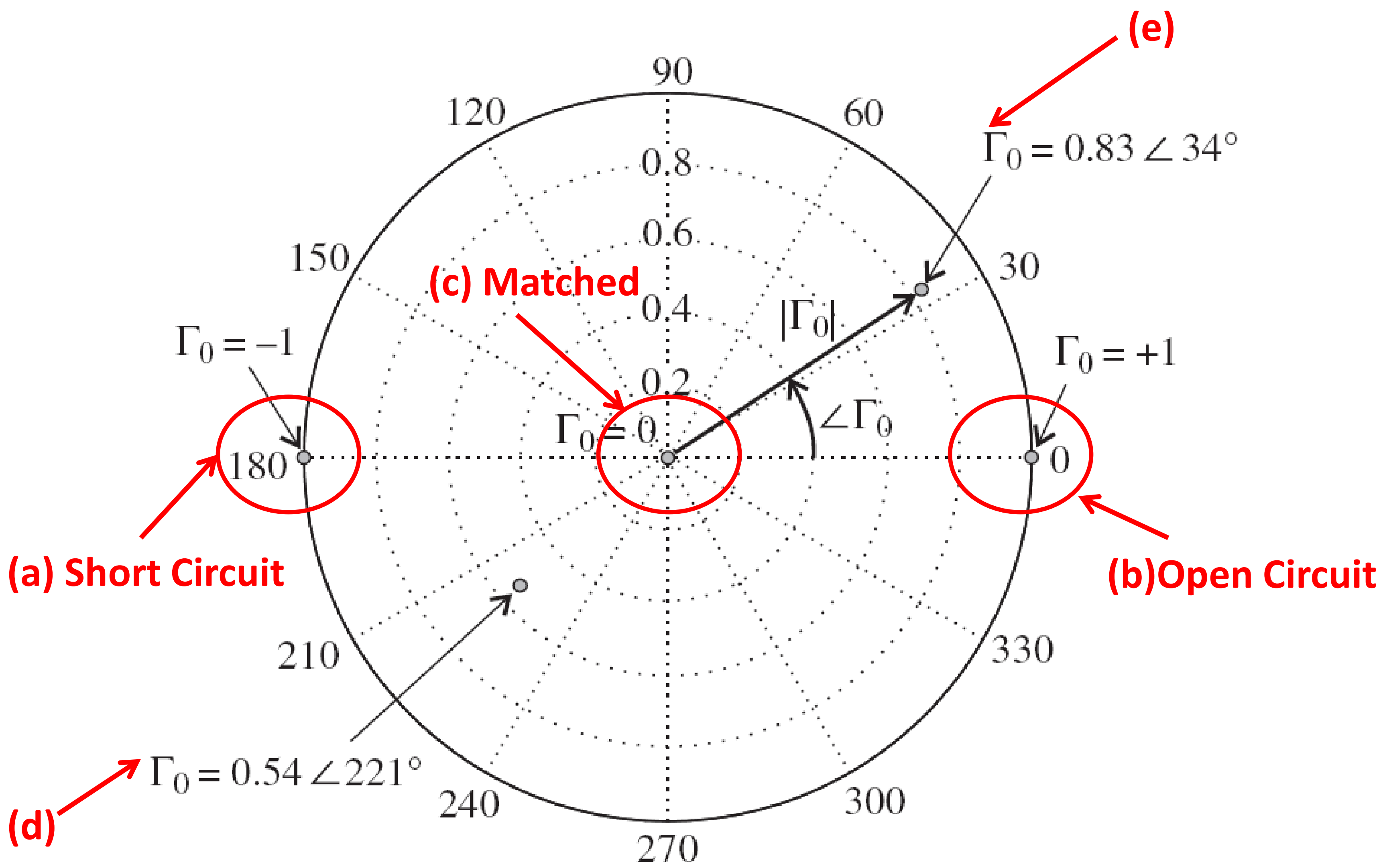
Display the respective reflection coefficients in complex Γ-plane

- **Solution:** We know the relationship between Z and Γ:

$$\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0} = \Gamma_{0r} + \Gamma_{0i} = |\Gamma_0| e^{j\theta_0}$$

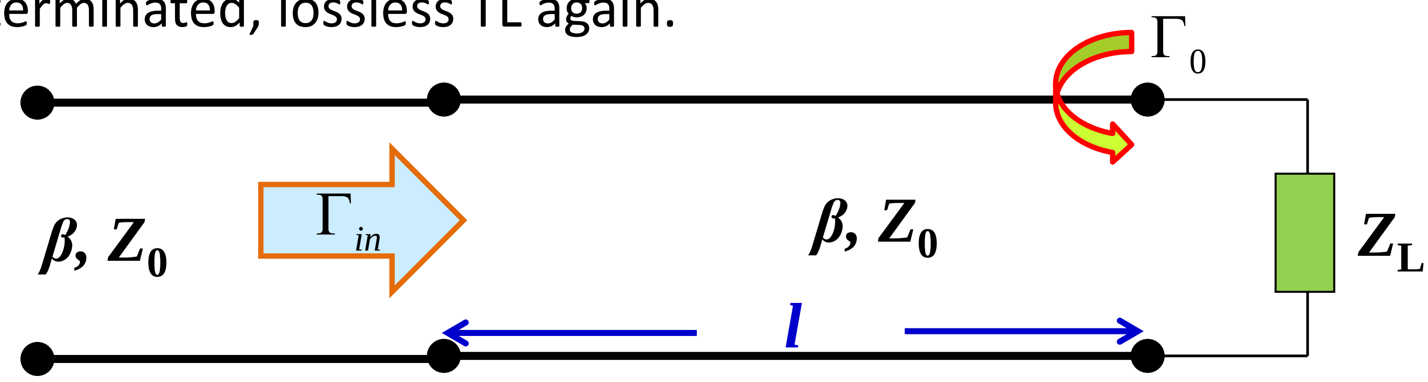
(a) Γ_0 = -1 (Short Circuit)
(b) Γ_0 = 1 (Open Circuit)
(c) Γ_0 = 0 (Matched)
(d) Γ_0 = 0.54<221°
(e) Γ_0 = 0.83<34°

Example – 1 (contd.)



Transformations on the Complex Γ-Plane

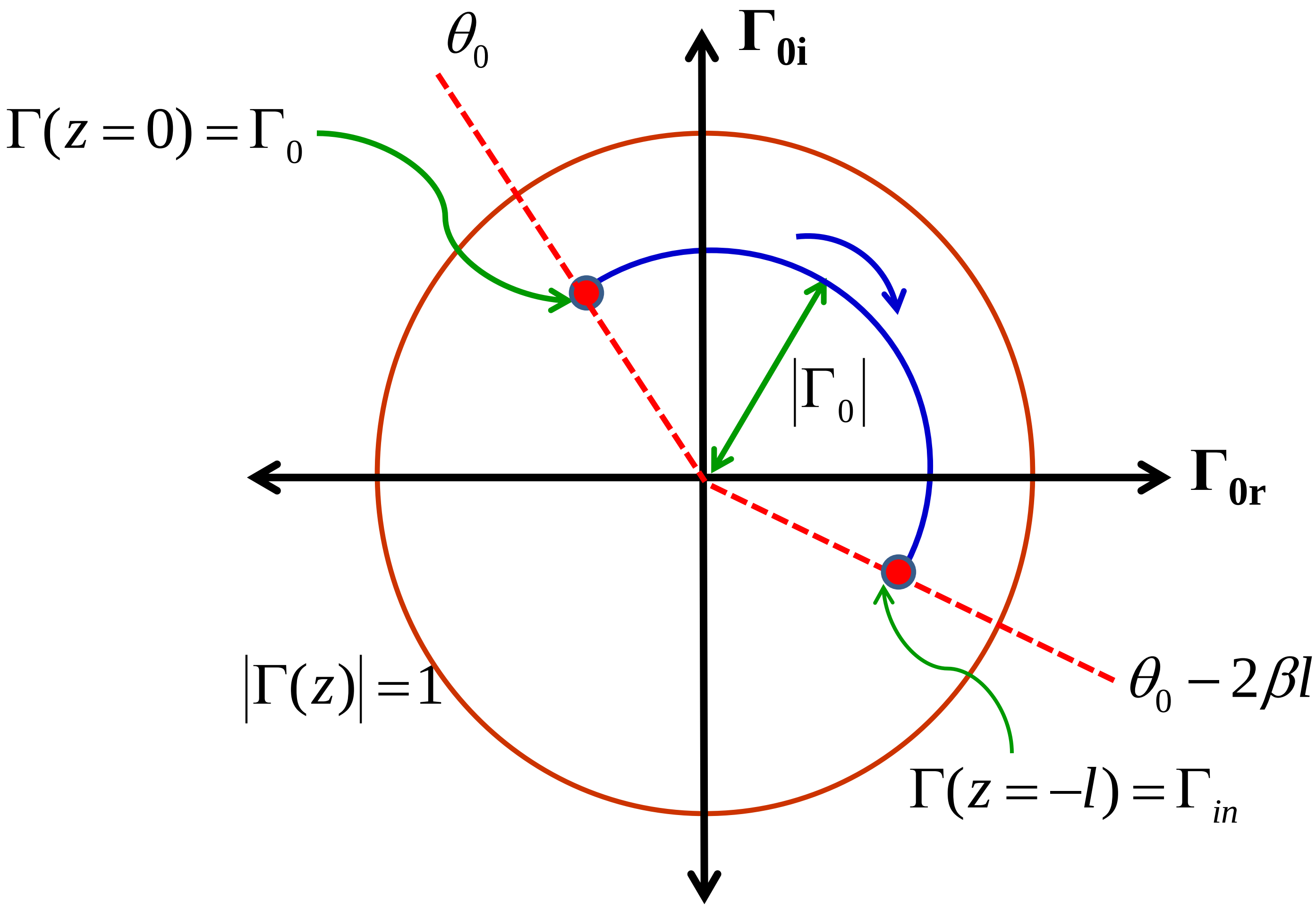
- The usefulness of the complex Γ-plane will be evident when we consider the terminated, lossless TL again.



- At $z = 0$, the reflection coefficient is called load reflection coefficient (Γ_0) → this actually describes the mismatch between the load impedance (Z_L) and the characteristic impedance (Z_0) of the TL.
- The move away from the load (or towards the input/source) in the negative z-direction (clockwise rotation) requires multiplication of Γ_0 by a factor $\exp(+j2\beta z)$ in order to explicitly define the mismatch at location 'z' known as $\Gamma(z)$.
- This transformation of Γ_0 to $\Gamma(z)$ is the key ingredient in Smith chart as a graphical design/display tool.

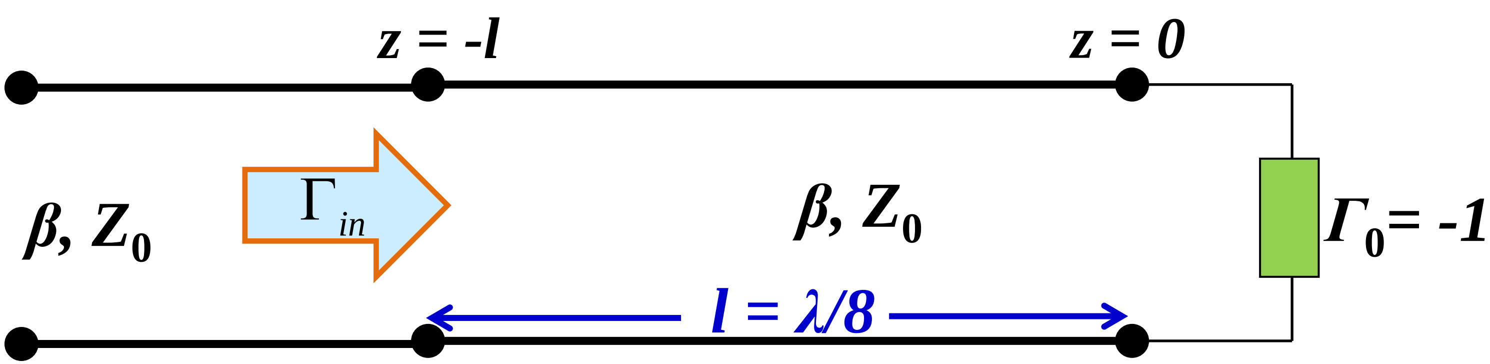
Transformations on the Complex Γ-Plane (contd.)

- Graphical interpretation of $\Gamma(z) = \Gamma_0 e^{+2j\beta z}$



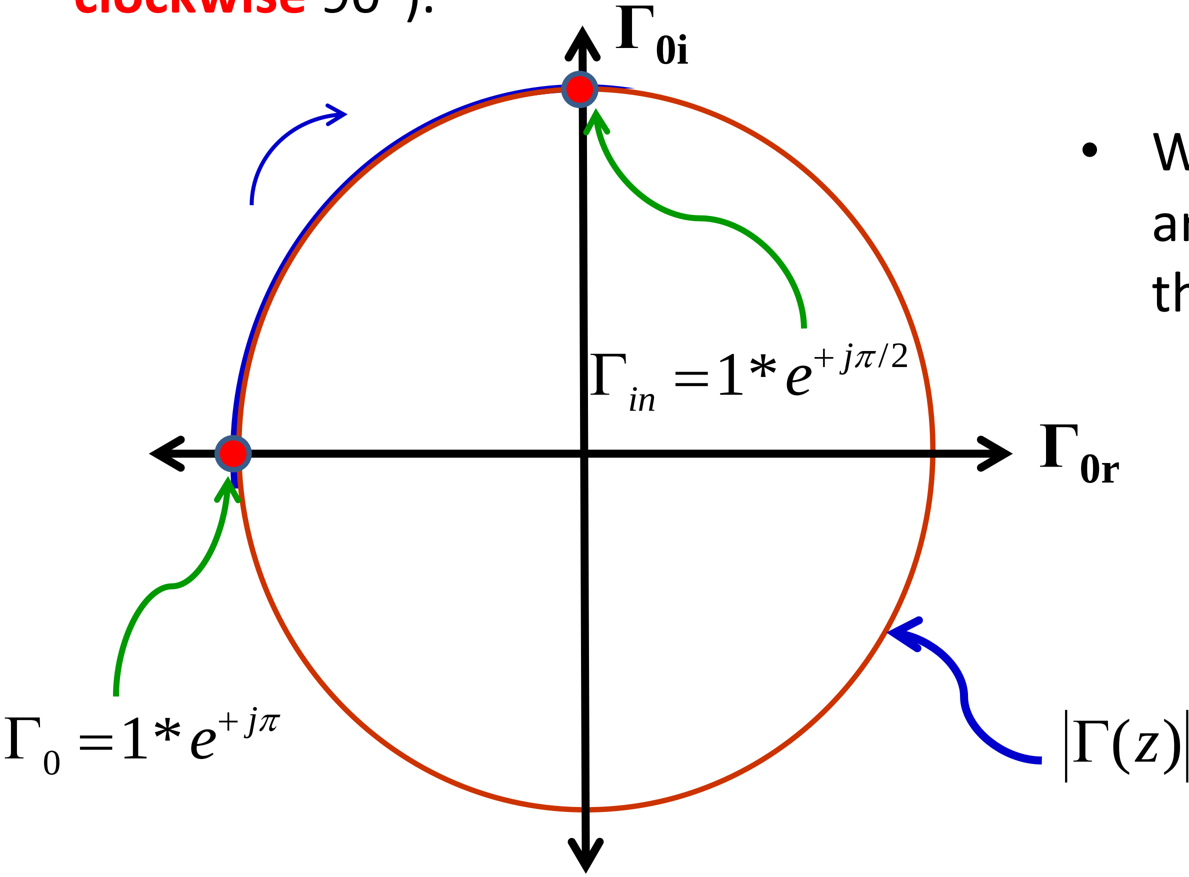
Transformations on the Complex Γ-Plane (contd.)

- It is clear from the graphical display that addition of a length of TL to a load Γ_0 **modifies** the **phase** θ_0 but **not** the **magnitude** Γ_0, we trace a **circular arc** as we parametrically plot $\Gamma(z)$! This arc has a **radius** Γ_0 and an **arc angle** $2\beta l$ radians.
- We can therefore **easily** solve many interesting TL problems **graphically**—using the complex Γ-plane! For **example**, say we wish to determine Γ_{in} for a transmission line length $l = \lambda/8$ and terminated with a **short** circuit.



Transformations on the Complex Γ-Plane (contd.)

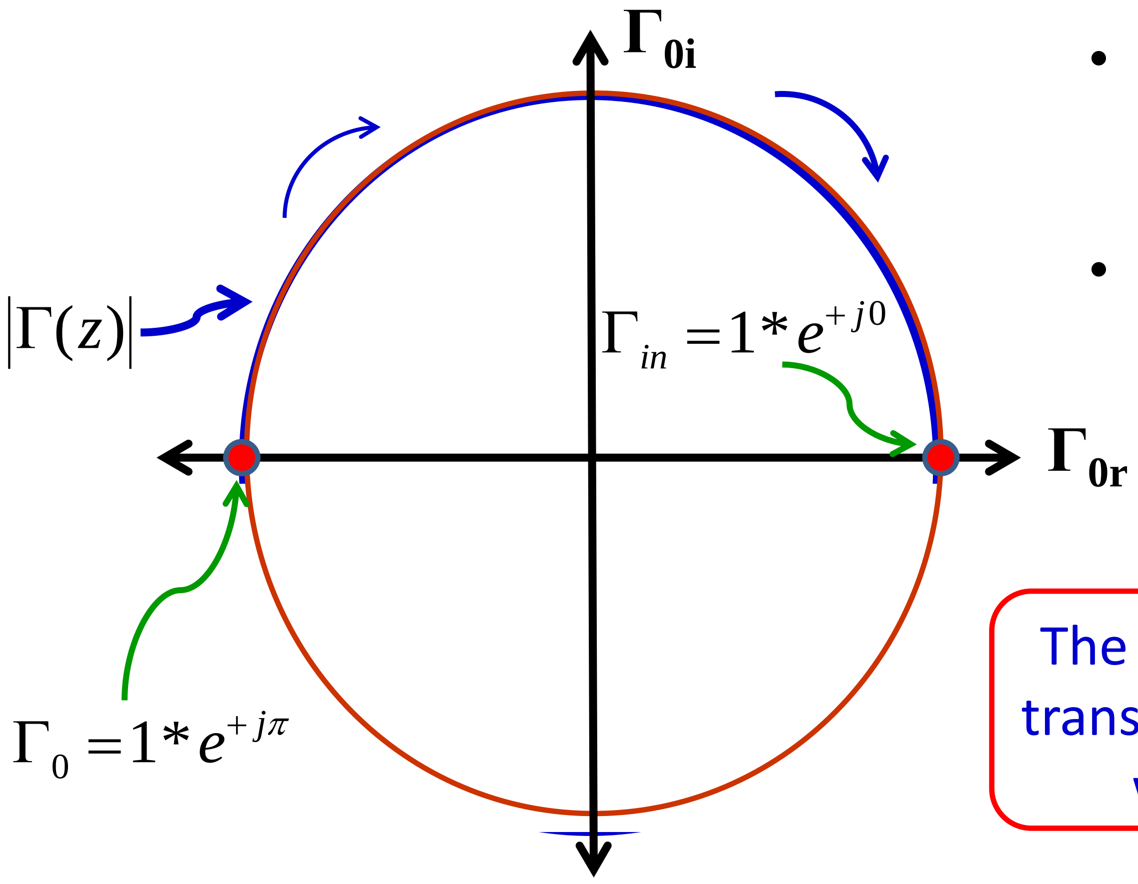
- The reflection coefficient of a **short** circuit is Γ_0 = −1 =1*e(jπ), and therefore we **begin** at the leftmost point on the complex Γ-plane. We then move along a **circular arc** $-2\beta l = -2(\pi/4) = -\pi/2$ radians (i.e., rotate **clockwise** 90^o).



- When we stop, we find we are at the point for Γ_{in}; in this case Γ_{in} = 1*e(jπ/2)

Transformations on the Complex Γ-Plane (contd.)

- Now let us consider the same problem, only with a new transmission line length $l = \lambda/4$.
- Now we rotate clockwise $2\beta l = \pi$ radians.

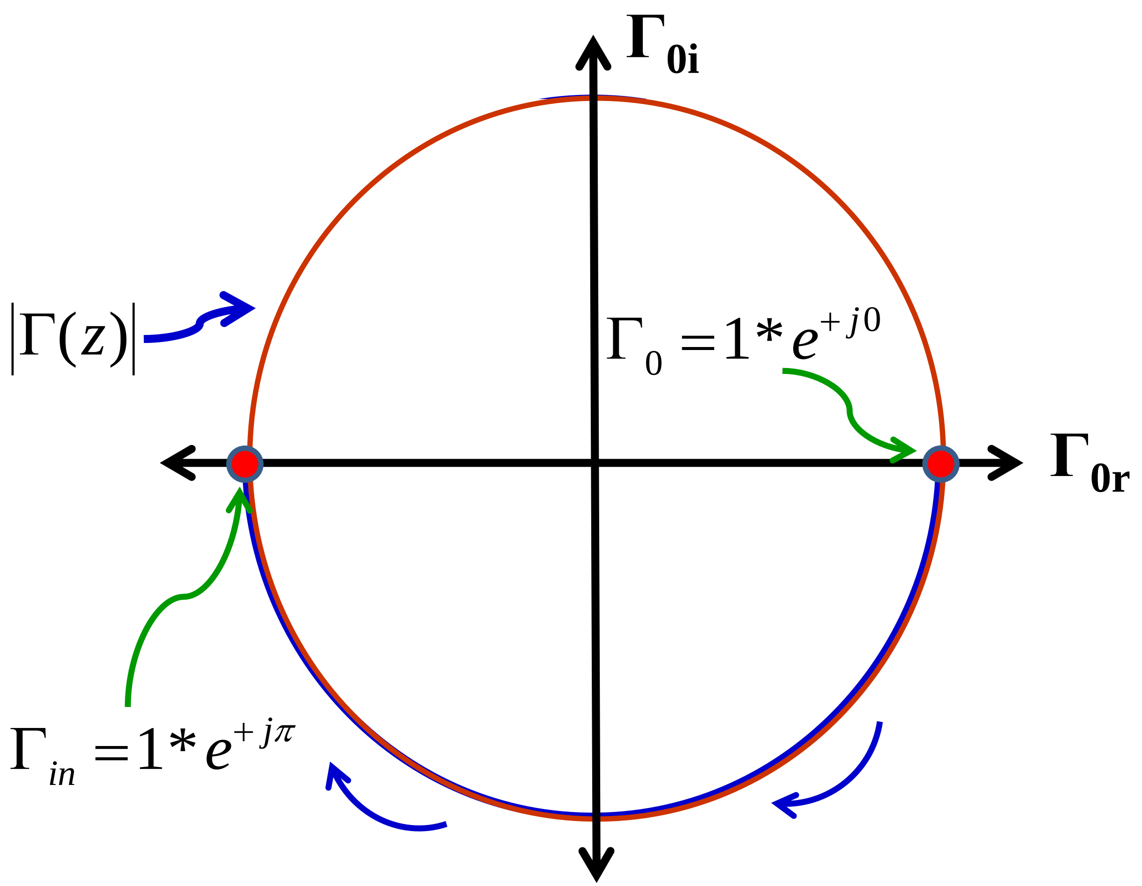


- In this case the input reflection coefficient is Γ_{in} = 1*e(j0) = 1
- The reflection coefficient of an open circuit

The short circuit load has been transformed into an open circuit with a quarter-wave TL

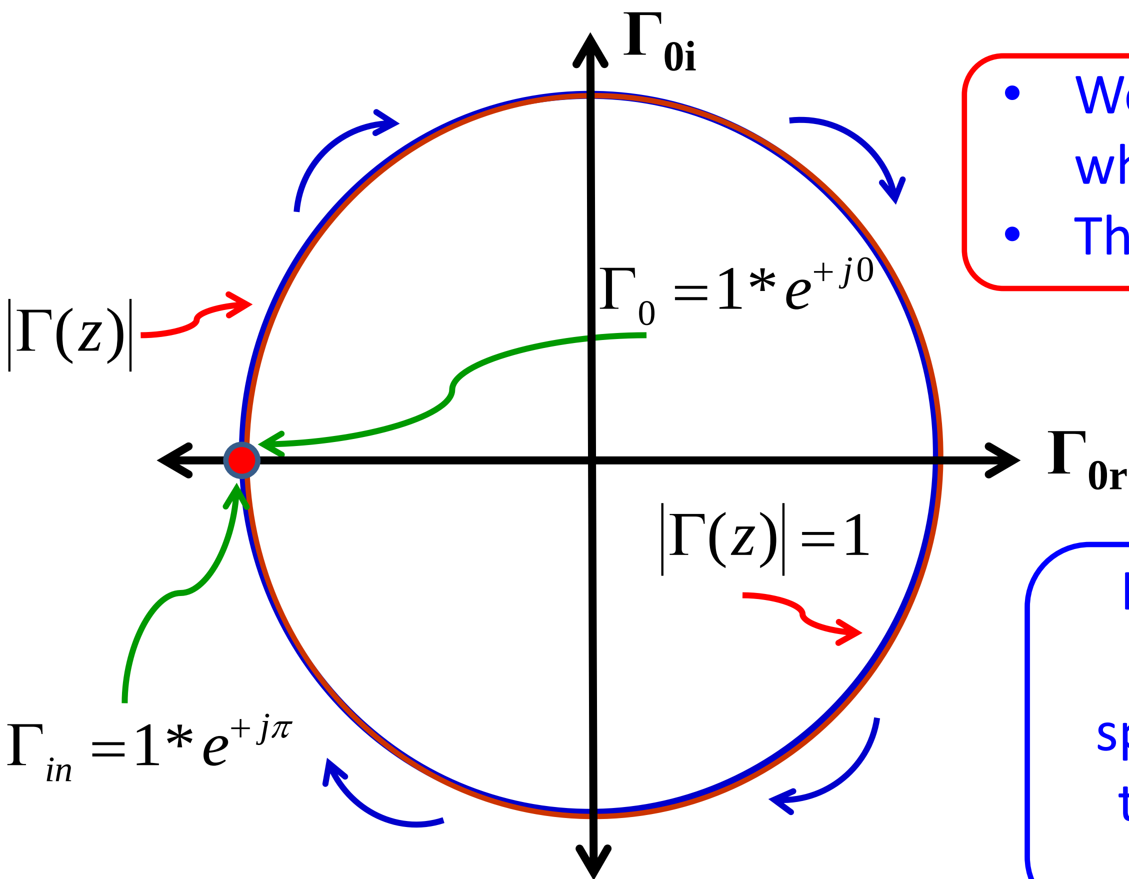
Transformations on the Complex Γ-Plane (contd.)

- We also know that a quarter-wave TL transforms an open-circuit into short-circuit → graphically it can be shown as:



Transformations on the Complex Γ-Plane (contd.)

- Now let us consider the same problem again, only with a new transmission line length $l = \lambda/2$.
- Now we rotate clockwise $2\beta l = 2\pi$ radians (360^o)

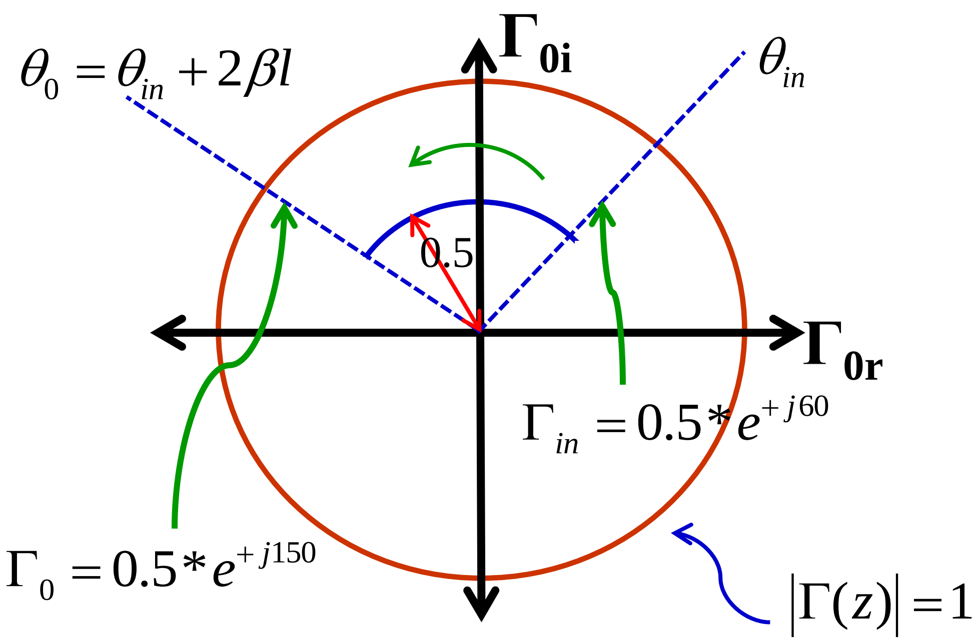


- We came clear around to where we started!
- Thus we conclude that $\Gamma_{in} = \Gamma_0$

It comes from the fact that **half-wavelength** TL is a special case, where we know that $Z_{in} = Z_L$ → eventually it leads to $\Gamma_{in} = \Gamma_0$

Transformations on the Complex Γ-Plane (contd.)

- Now let us consider the **opposite** problem. Say we know that the **input** reflection coefficient at the beginning of a TL with length l = λ/8 is: $\Gamma_{in} = 0.5e(j60°)$.
- What is the reflection coefficient at the **load**?
- In this case we rotate counter-clockwise along a circular arc (radius =0.5) by an amount $2\beta l = \pi/2$ radians (90^o).
- In essence, we are removing the phase associated with the TL.



The reflection coefficient at the load is:

$$\Gamma_0 = 0.5 * e^{+j150}$$

Mapping Z to Γ

- We know that the line impedance and reflection coefficient are **equivalent** – either one can be expressed in terms of the other.

$$\Gamma(z) = \frac{Z(z) - Z_0}{Z(z) + Z_0} \quad \Longleftrightarrow \quad Z(z) = Z_0\left(\frac{1+\Gamma(z)}{1-\Gamma(z)}\right)$$

- The above expressions depend on the characteristic impedance Z_0 of the TL. In order to generalize the relationship, we first define a **normalized** impedance value **z'** as:

$$z'(z) = \frac{Z(z)}{Z_0} = \frac{R(z)}{Z_0} + j\frac{X(z)}{Z_0} = r(z) + jx(z)$$

therefore

$$\Gamma(z) = \frac{Z(z) - Z_0}{Z(z) + Z_0} = \frac{\left(Z(z)/Z_0\right) - 1}{\left(Z(z)/Z_0\right) + 1} = \frac{z'(z) - 1}{z'(z) + 1}$$

$$z'(z) = \frac{1+\Gamma(z)}{1-\Gamma(z)}$$

Mapping Z to Γ (contd.)

$$\Gamma(z)=\frac{Z(z)-Z_0}{Z(z)+Z_0}=\frac{\left(Z(z)/Z_0\right)-1}{\left(Z(z)/Z_0\right)+1}=\frac{z'(z)-1}{z'(z)+1}$$

$$z'(z)=\frac{1+\Gamma(z)}{1-\Gamma(z)}$$

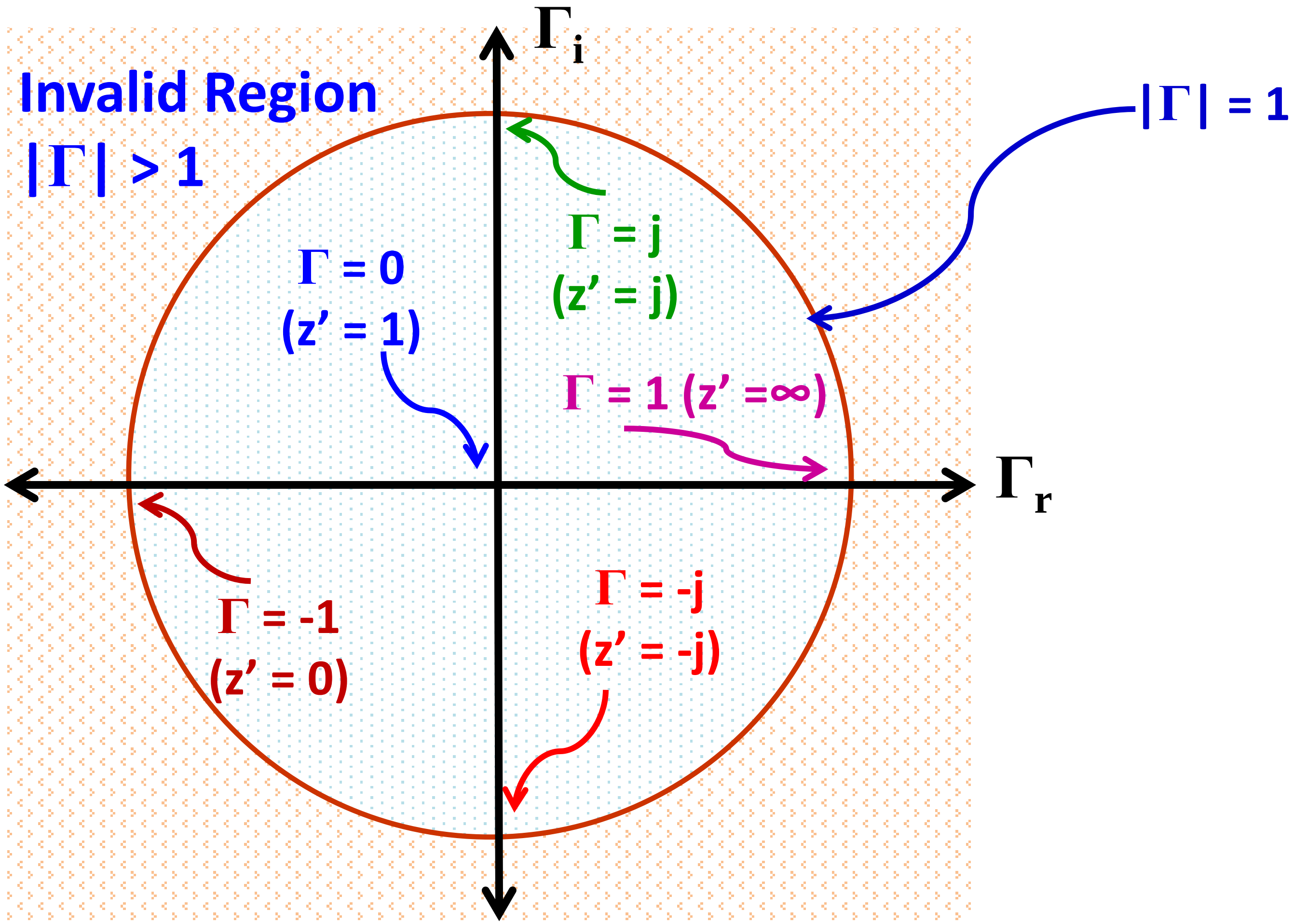
These equations describe a mapping between z' and Γ. That means that each and every normalized impedance value likewise corresponds to one specific point on the complex Γ-plane

- For example, we wish to indicate the values of some common normalized impedances (shown below) on the complex Γ-plane and vice-versa.

Case	Z	z'	Γ
1	∞	∞	1
2	0	0	-1
3	Z_0	1	0
4	jZ_0	j	j
5	$-jZ_0$	-j	-j

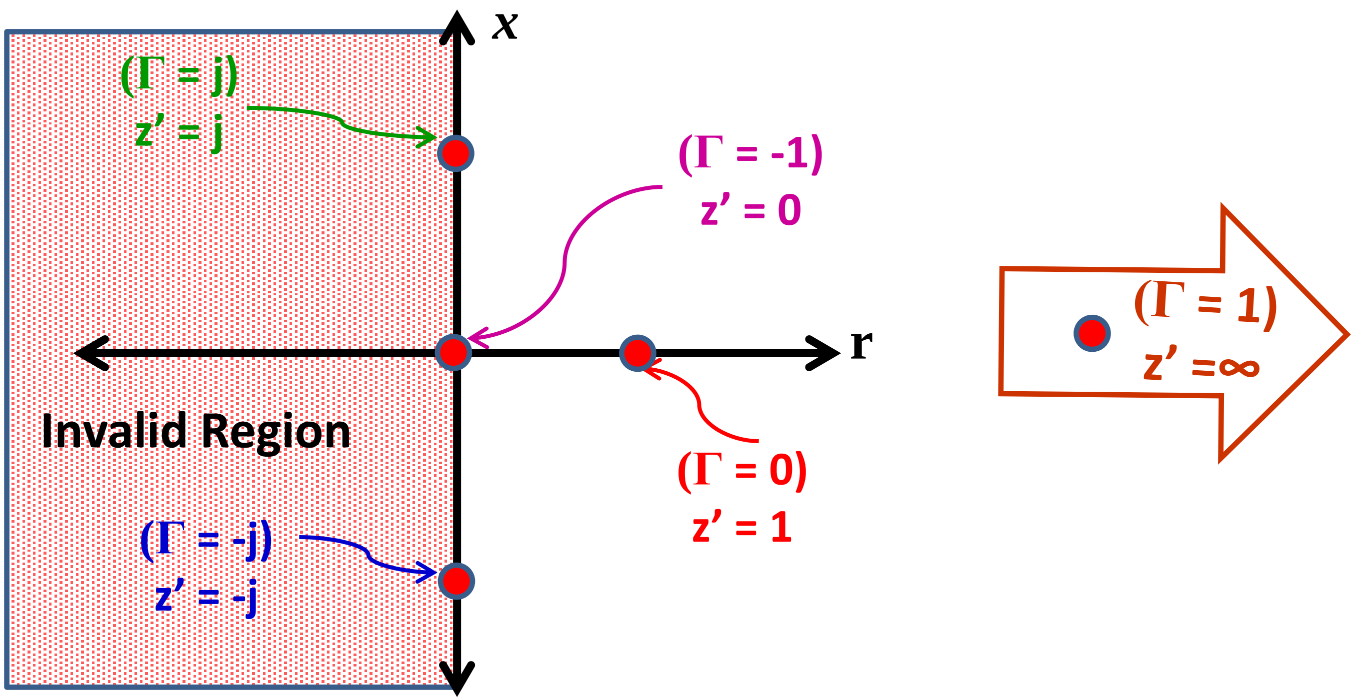
Mapping Z to Γ (contd.)

- The five normalized impedances map five specific points on the complex Γ-plane.



Mapping Z to Γ (contd.)

- The five complex-Γ map onto five points on the normalized Z-plane

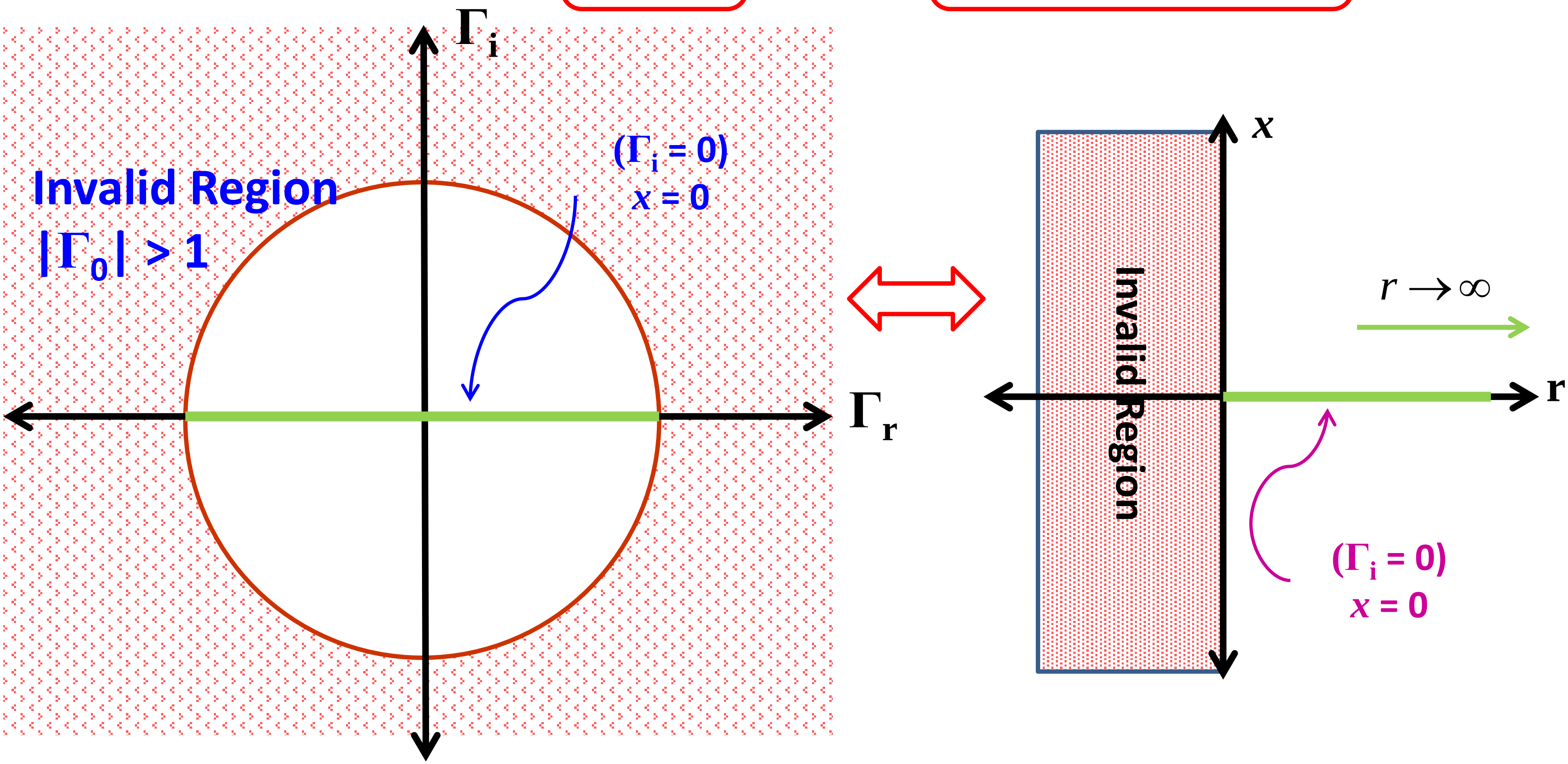


- It is apparent that the normalized impedances can be mapped on complex Γ-plane and vice versa
- It gives us a clue that whole impedance contours (i.e, set of points) can be mapped to complex Γ-plane

Mapping Z to Γ (contd.)

Case-I: Z = R → impedance is purely real

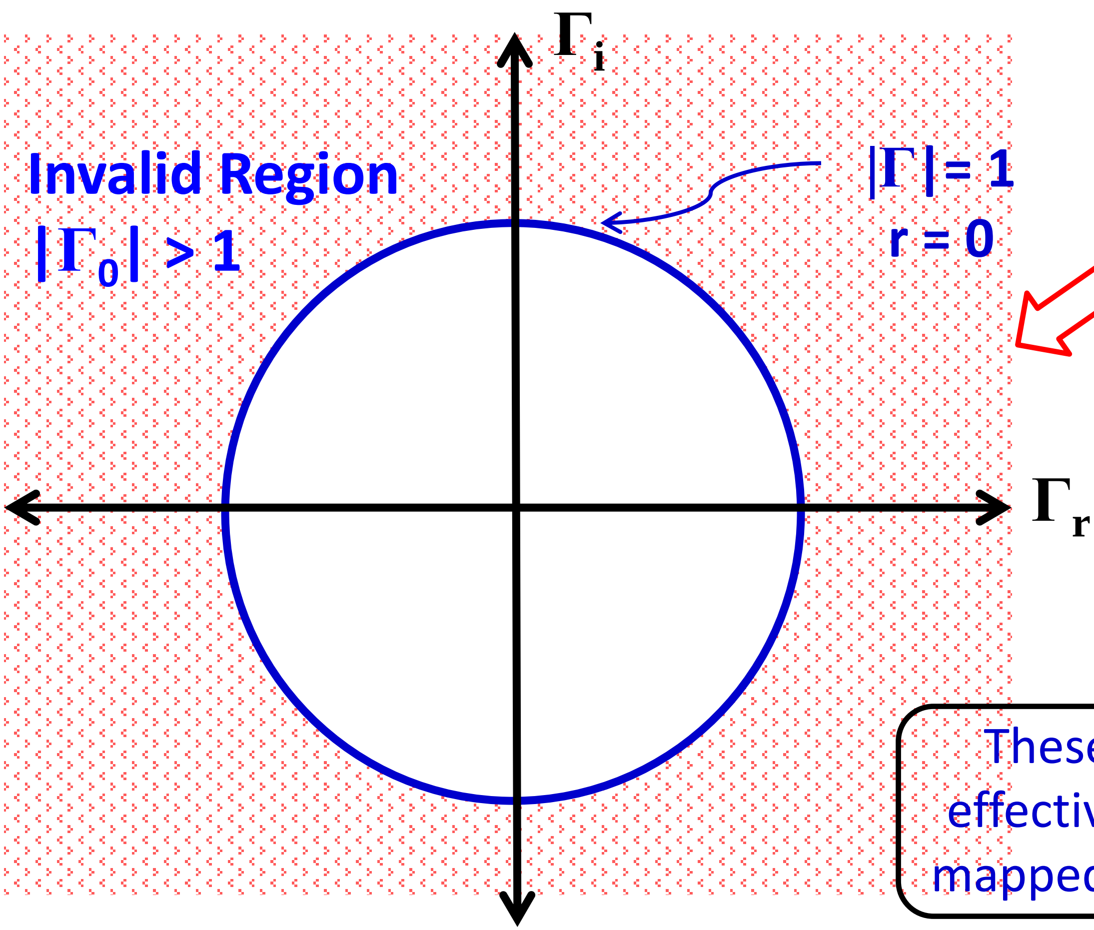
$$z' = r + j0 \quad \Rightarrow \quad \Gamma = \frac{r-1}{r+1} \quad \Rightarrow \quad \Gamma_r = \frac{r-1}{r+1} \qquad \Gamma_i = 0$$

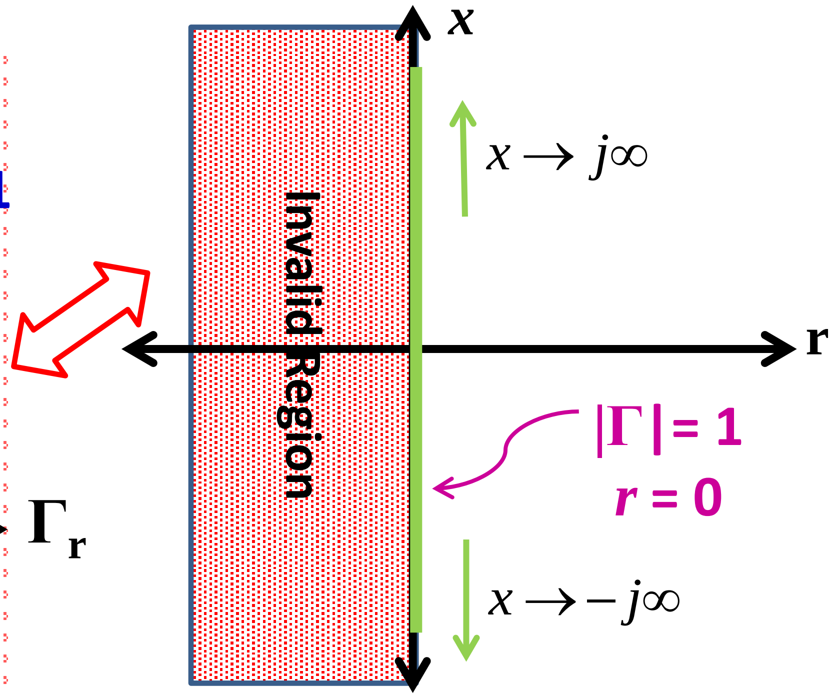


Mapping Z to Γ (contd.)

Case-II: Z = jX → impedance is purely imaginary

$z' = 0 + jx$ ← Purely reactive impedance results in a reflection coefficient with unity magnitude $|\Gamma| = 1$





These cases (I and II) demonstrate that effectively any complex impedance can be mapped to complex Γ-plane → **Smith Chart**

The Smith Chart

In summary

- A vertical line $r = 0$ on complex Z-plane maps to a circle $|\Gamma| = 1$ on the complex Γ-plane
- A horizontal line $x = 0$ on complex Z-plane maps to the line $\Gamma_i = 0$ on the complex Γ-plane

Very fascinating in an academic sense, but are not relevant considering that actual values of impedance generally have both a real and imaginary component

Mappings of more general impedance contours (e.g, $r = 0.5$ and $x = -1.5$ corresponding to normalized impedance $0.5 - j1.5$) can also be mapped

Smith Chart

The Smith Chart (contd.)

- Let us revisit the generalized reflection coefficient formulation:

$$\Gamma(z) = |\Gamma_0| e^{j\theta_0} e^{j2\beta z} = \Gamma_r + j\Gamma_i$$

- Therefore, the normalized impedance can be formulated as:

$$z'(z) = r + jx = \frac{Z(z)}{Z_0} = \frac{1+\Gamma(z)}{1-\Gamma(z)} = \frac{1+\Gamma_r + j\Gamma_i}{1-\Gamma_r - j\Gamma_i}$$

$$\Rightarrow \left((1-\Gamma_r) - j\Gamma_i\right)(r + jx) = (1+\Gamma_r) + j\Gamma_i$$

- The separation of real and imaginary part results in:

$$r(1-\Gamma_r) + x\Gamma_i = (1+\Gamma_r)$$ ← **Real**

$$x(1-\Gamma_r) - r\Gamma_i = \Gamma_i$$ ← **Imaginary**

The Smith Chart (contd.)

- Simplification and then elimination of reactance (x) from these two give:

$$(1-\Gamma_r)r+\Gamma_i\left[\left(\frac{\Gamma_i}{1-\Gamma_r}\right)(1+r)\right]=1+\Gamma_r$$

Multiplying through by $1-\Gamma_r$

$$(1-\Gamma_r)^2 r+\Gamma_i^2(1+r)=(1+\Gamma_r)(1-\Gamma_r)$$

$$(1-\Gamma_r)^2 r+\Gamma_i^2(1+r)=1-\Gamma_r^2$$

$$\Gamma_r^2(1+r)-2\Gamma_r r+(r-1)+\Gamma_i^2(1+r)=0$$

The Smith Chart (contd.)

$$\Gamma_r^2(1+r) - 2\Gamma_r r + \Gamma_i^2(1+r) = 1 - r$$

$$\Gamma_r^2 - 2\Gamma_r\left(\frac{r}{1+r}\right) + \Gamma_i^2 = \frac{1-r}{1+r}$$

$$\left(\Gamma_\mathrm{r} - \frac{\mathrm{r}}{1+\mathrm{r}}\right)^2 + \Gamma_\mathrm{i}^2 = \frac{1-\mathrm{r}}{1+\mathrm{r}} + \left(\frac{\mathrm{r}}{1+\mathrm{r}}\right)^2$$

$$\left(\Gamma_r - \frac{r}{1+r}\right)^2 + \Gamma_i^2 = \frac{(1+r)(1-r) + (r)^2}{(1+r)^2}$$

The Smith Chart (contd.)

$$\left(\Gamma_r - \frac{r}{1+r}\right)^2 + \Gamma_i^2 = \frac{1}{(1+r)^2}$$

Similar equation to circle of radius l, centered at (p,q):

$$(\Gamma_r - p)^2 + (\Gamma_i - q)^2 = l^2$$

This is equation of a circle

center: $(p,q) = \left(\frac{r}{1+r}, 0\right)$ **and** **radius:** $l = \frac{1}{1+r}$

Observations:

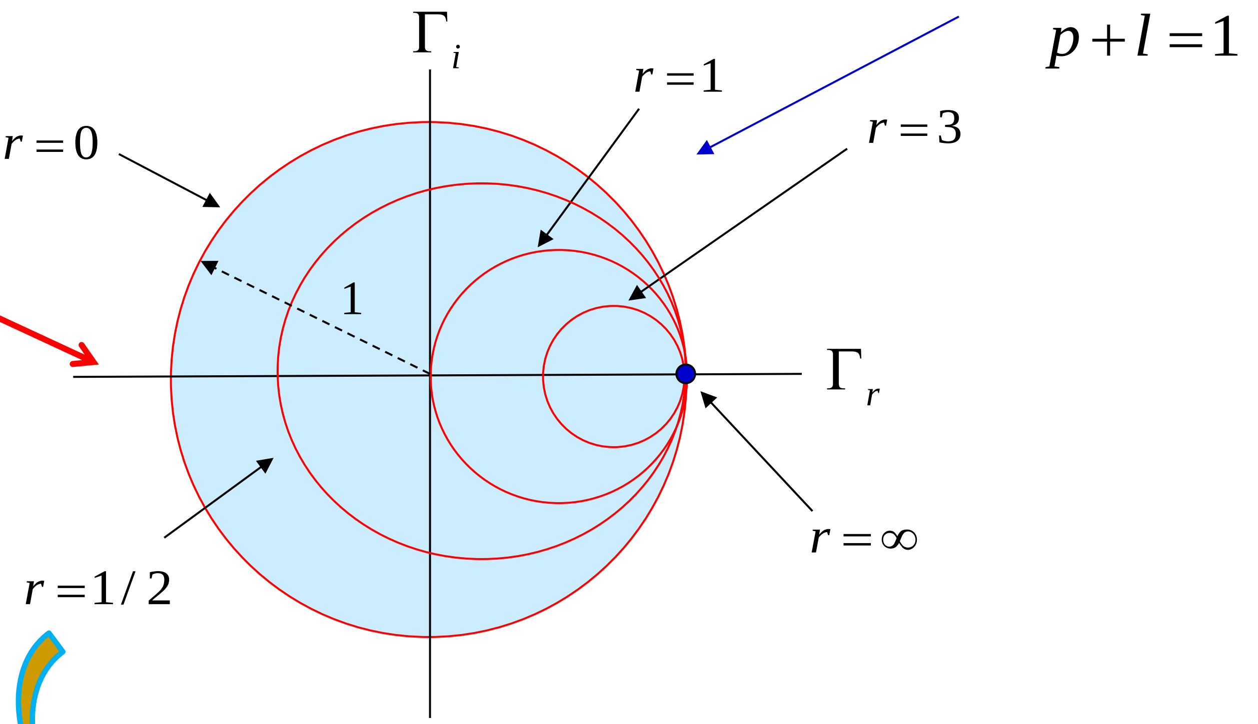
- **For r =0:** $p^2 + q^2 = 1$; (p, q) = (0, 0) and $l = 1$
- **For r =1/2:** $(p - 1/3)^2 + q^2 = (2/3)^2$; (p, q) = (1/3, 0) and $l = 2/3$
- **For r =1:** $(p - 1/2)^2 + q^2 = (1/2)^2$; (p, q) = (1/2, 0) and $l = 1/2$
- **For r =3:** $(p - 3/4)^2 + q^2 = (1/4)^2$; (p, q) = (3/4, 0) and $l = 1/4$

Circles of distinct centre and radii

The Smith Chart (contd.)

Therefore the resistance circles on the complex Γ-plane are:

Because of $(q-0)^2$ term, all the constant resistance (r) circles have centers on this line



This approach enables mapping of any realizable vertical line (representing **r**) in the complex Γ-plane

The Smith Chart (contd.)

- For the mapping of horizontal lines of the normalized impedance plane to Γ-plane, let us simplify and eliminate resistance (*r*) from the following:

$$r(1-\Gamma_r)+x\Gamma_i=(1+\Gamma_r)$$ ← Real

$$x(1-\Gamma_r)-r\Gamma_i=\Gamma_i$$ ← Imaginary

$$(1-\Gamma_r)\left[\frac{(1-\Gamma_r)x-\Gamma_i}{\Gamma_i}\right]+x\Gamma_i=1+\Gamma_r$$

$$(1-\Gamma_r)^2x-\Gamma_i(1-\Gamma_r)+x\Gamma_i^2-\Gamma_i(1+\Gamma_r)=0$$

$$(1-\Gamma_r)^2x-2\Gamma_i+x\Gamma_i^2=0$$

The Smith Chart (contd.)

$$(\Gamma_r - 1)^2 - \left(\frac{2}{x}\right)\Gamma_i + \Gamma_i^2 = 0$$

$$(\Gamma_r - 1)^2 + \left(\Gamma_i - \frac{1}{x}\right)^2 = \left(\frac{1}{x}\right)^2$$

center: $(p, q) = (1, 1/x)$

radius: $l = \frac{1}{|x|}$

Note:

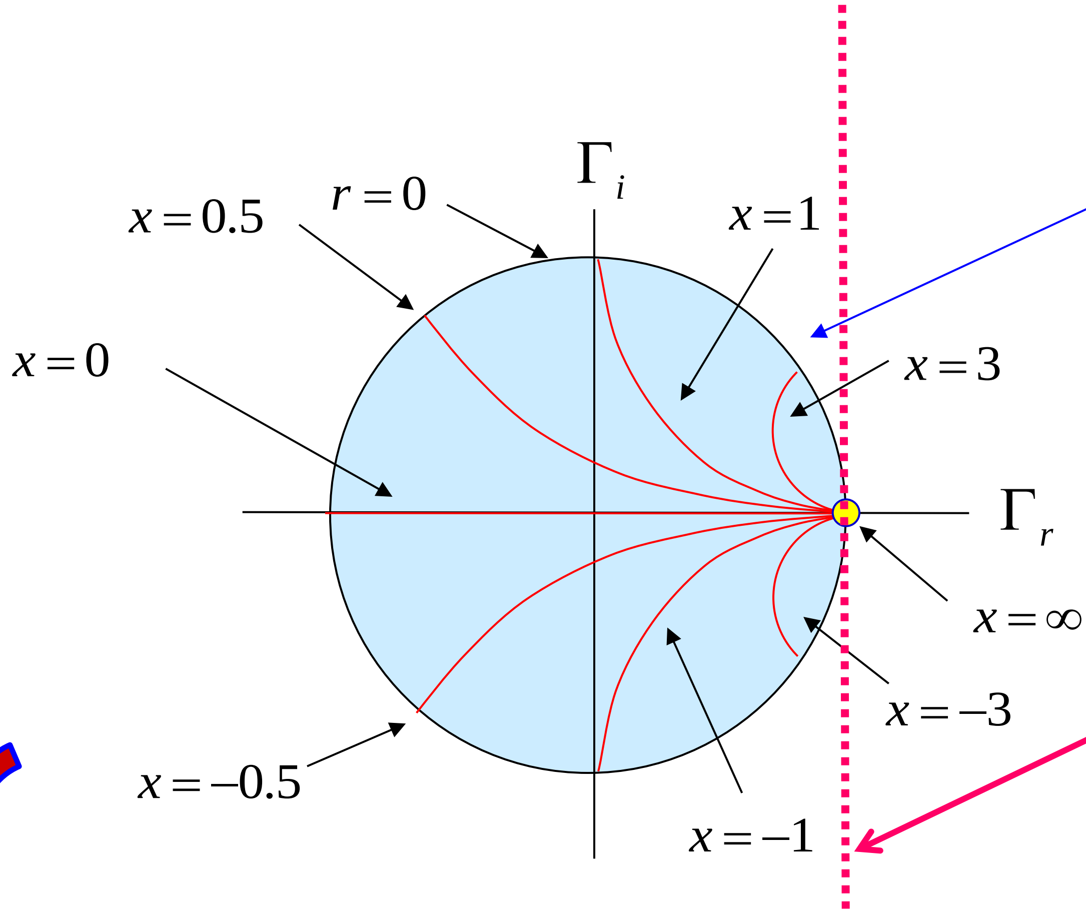
$q = \pm l$

Observations:

- **For x =1:** $(p – 1)^2 + (q – 1)^2 = (1)^2$; (p, q) = (1, 1) and l = 1
- **For x =-1:** $(p – 1)^2 + (q + 1)^2 = (1)^2$; (p, q) = (1, -1) and l = 1
- **For x =1/2:** $(p – 1)^2 + (q – 2)^2 = (2)^2$; (p, q) = (1, 2) and l = 2
- **For x =-1/2:** $(p – 1)^2 + (q + 2)^2 = (2)^2$; (p, q) = (1, -2) and l = 2

Circles of distinct centre and radii

The Smith Chart (contd.)



Note:

$q = \pm l$

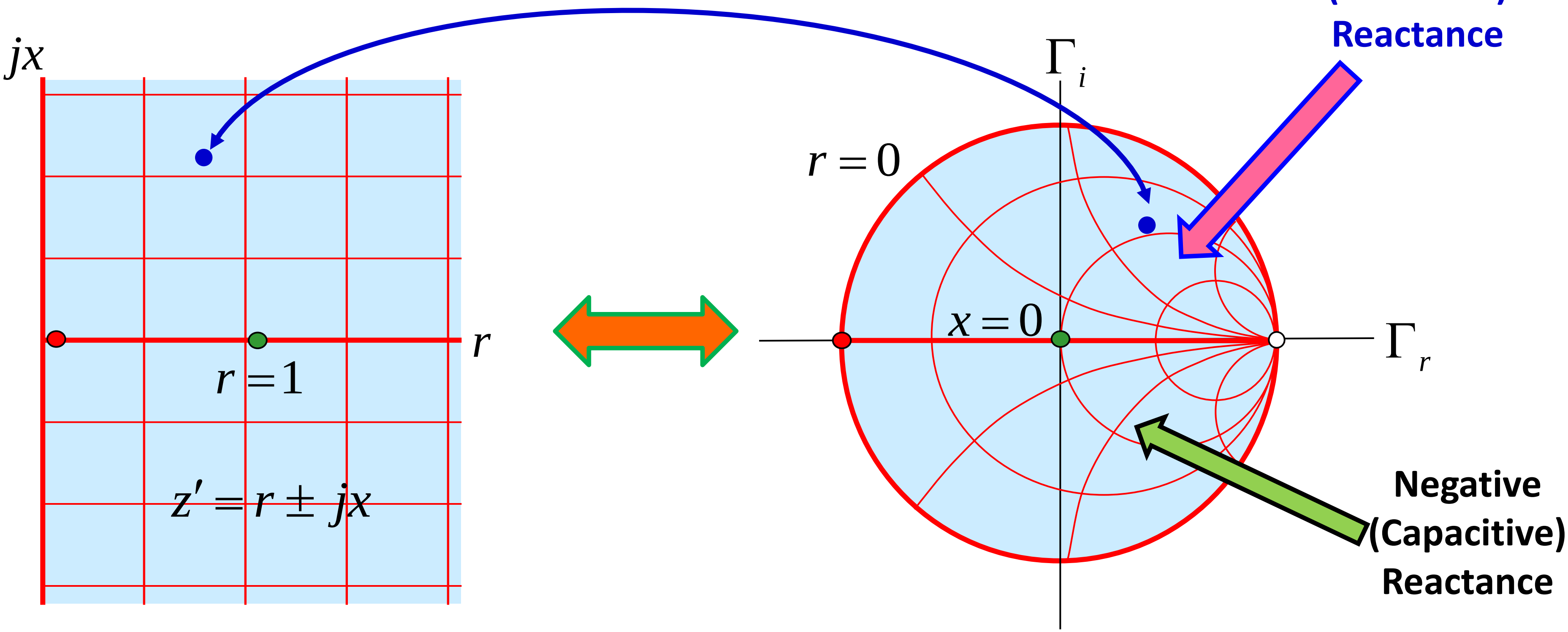
All constant reactance (x) circles have their origins along this line p=1 because of the term (p – 1)²

This approach enables mapping of any realizable horizontal line (representing x) in the complex Γ-plane

The Smith Chart (contd.)

- Combination of these constant resistance and reactance circles define the mappings from normalized impedance (z’) plane to Γ-plane and is called as Smith chart.

$$z'(z) = \frac{1+\Gamma(z)}{1-\Gamma(z)} \quad \Leftrightarrow \quad \Gamma(z) = \frac{z'(z)-1}{z'(z)+1}$$



The Smith Chart (contd.) – Important Features

1. By definition:

$$\Gamma(z) = \frac{z'(z)-1}{z'(z)+1} = \frac{r+jx-1}{r+jx+1} \quad \Rightarrow \quad |\Gamma(z)| = \frac{(r-1)^2+x^2}{(r+1)^2+x^2}$$

- It is apparent: for $r \geq 0$, we get $|\Gamma(z)| \leq 1$. This condition is easily met for passive networks (i.e, no amplifiers) and lossless TLs (real Z_0)
- Consequently, the standard Smith chart only shows only the inside of the unit circle in the Γ-plane. That is, $|\Gamma(z)| \leq 1$ which is bounded by the $r = 0$ circle described by:

$$\Gamma_r^2 + \Gamma_i^2 = 1$$

The Smith Chart (contd.) – Important Features

2. Notice that in the upper semi-circle of the Smith chart, $x \geq 0$ which is an inductive reactance. Consequently, the generalized reflection coefficients $\Gamma(z) \equiv \Gamma_r + j\Gamma_i$ in the upper semi-circle are associated with normalized TL impedances $z'(z) \equiv r + jx$ that are inductively reactive.

   Conversely, the lower semi-circle of the Smith chart represent capacitive reactive impedances

3. If z'(z) is purely real (ie, $x = 0$) then the reactance term:

$$\left(1-\Gamma_r\right)^2 x - 2\Gamma_i + x\Gamma_i^2 = 0$$

→ suggests → **$\Gamma_i = 0$ except possibly at $\Gamma_r = 1$**

   Consequently, purely real z'(z) values are mapped to $\Gamma(z)$ values on the Γ_r axis.

The Smith Chart (contd.) – Important Features

4. If z’(z) is purely imaginary (ie, r= 0) then the impedance term:

$$\left(\Gamma_r - \frac{r}{1+r}\right)^2 + \Gamma_i^2 = \frac{(1+r)(1-r)+(r)^2}{(1+r)^2}$$

suggests

$$\Gamma_r^2 + \Gamma_i^2 = 1$$

Unit Circle on Γ-plane

Consequently, purely imaginary z’(z) values are mapped to $\Gamma(z)$ values on the unit circle in Γ-plane.