CHAPTER

21 Radar

The idea that radio waves could be used to detect the presence of stationary or moving objects emerged around 1900, almost as soon as radio itself. Christian Hueslmeyer, a German inventor, demonstrated an apparatus in 1904 which, when mounted on a bridge above the Rhine, rang a bell when a ship passed beneath. He used a (now) primitive spark gap RF source and coherer detector. The system may have shown only marginal potential for collision avoidance, as the German Navy demonstrated no interest. Sir Robert Watson-Watt developed meteorological radar in Britain in the 1930s and then a chain of air defense radars during World War II. In the U.S., the MIT Radiation Laboratory, set up to develop military microwave radar systems, had nearly 4000 employees between 1940 and 1945. The acronym RADAR, for Radio Detection And Ranging, has been attributed to U.S. Navy officers F. R. Furth and S. M. Tucker, who introduced it in 1940, though the term remained classified throughout the war.[1]

Today, radar goes beyond aircraft tracking to applications as diverse as space object monitoring, storm tracking, detection of clear air turbulence, and velocity measurements of speeding automobiles and baseballs. In this chapter, we look at some commonly used radars, some general system aspects of radar, and, finally, some RF components and techniques developed specifically for radar.

21.1 Some representative radar systems

Classic surveillance radar

Figure 21.1 is a block diagram of the classic radar system used to monitor air traffic. A rotating antenna sweeps continuously in azimuth while it repeatedly transmits short pulses and receives subsequent echos from targets. As each pulse is transmitted, the beam of the CRT display begins to sweep from the center in a radial direction. The detected signal from the receiver modulates the

[1] Butrica, A. J. *To See the Unseen: A History of Planetary Radar Astronomy*, Diane Publishing Co. 1997 (also available on the Web).

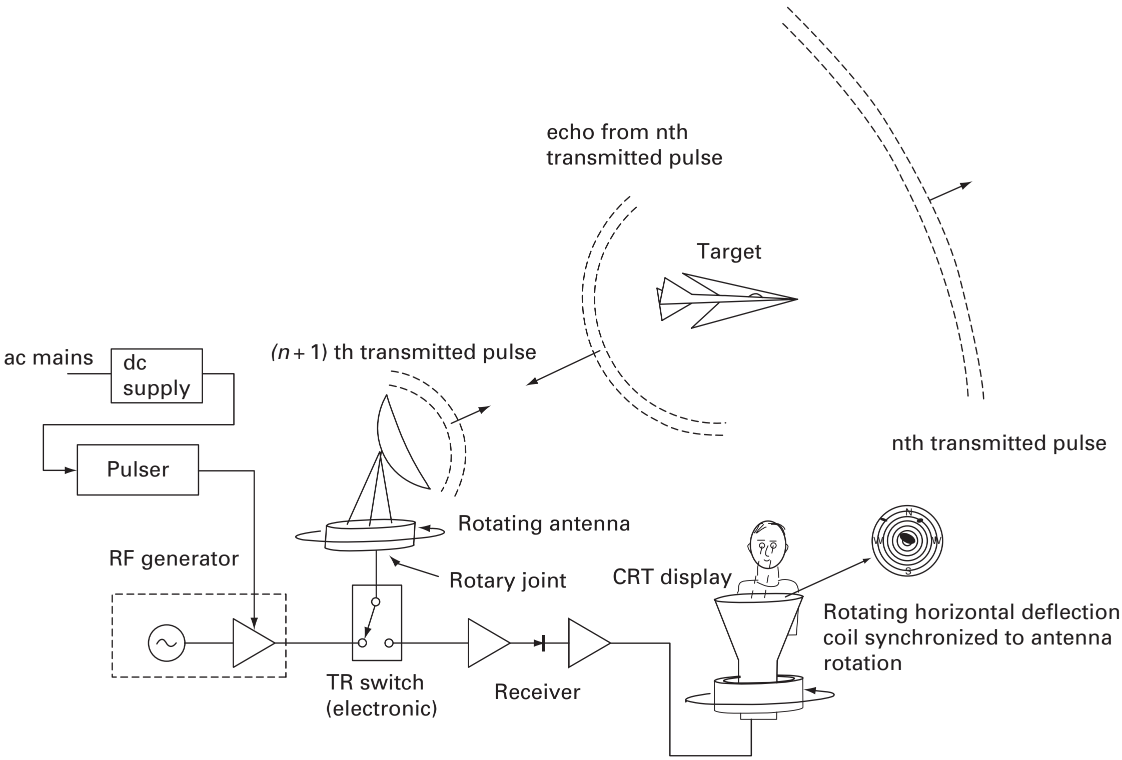

Figure 21.1. Classic surveillance radar.

intensity of the spot, rather than causing a deflection perpendicular to the sweep direction. The beam deflection coil is made to rotate continuously around the neck of the CRT, synchronized to the rotation of the antenna, so that the radial sweep direction at any instant is the azimuthal pointing direction of the antenna.

Generally there is a bright area near the center of the CRT, produced by reflections from nearby hills, buildings and towers. The receiver may incorporate an electronic gain control, slaved to the CRT sweep to compensate for the decrease in echo strength with increasing target range. Dashed lines on the figure show the spherical wavefronts for two transmitted pulses and for the reflection (echo) from one pulse, but do not indicate that the intensity of the transmitted waves is sharply concentrated in the forward direction by the high directivity of the antenna.

21.1.1 CW Doppler radar

The classic radar described above uses short pulses for good range resolution. It detects the amplitude of each returning pulse, paying no attention to the phase. Quite the opposite is the radar speed gun, which transmits cw and therefore has no range resolution, but does detect the phase to measure speed. A functional block diagram of the speed gun is shown in Figure 21.2.

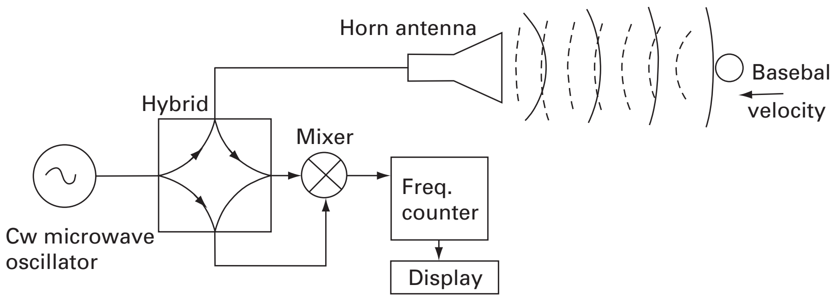

Figure 21.2 Radar speed gun.

The signal from a cw microwave oscillator in the 10–20 GHz range (usually a Gunn diode negative resistance oscillator) is split to feed the horn antenna and a mixer (multiplier). The echo from the target also provides a signal to the mixer. A moving target causes a Doppler shift, so the echo has a higher frequency if the target is approaching the radar. The output from the mixer contains a signal whose frequency is equal to the Doppler shift. This frequency is counted and multiplied by the appropriate factor so that the display reads line-of-sight velocity. Note that the Doppler shift depends only on the line-of-sight velocity of the target. This radar is totally insensitive to range.

21.1.2 Chirped cw radar

The radar in Figure 21.3 differs from the radar speed gun only in that its oscillator is a "chirped" VCO, whose output frequency follows a sawtooth control voltage. Here this radar is used to determine the position of a stationary (or very nearly stationary) fluid level in a storage tank.

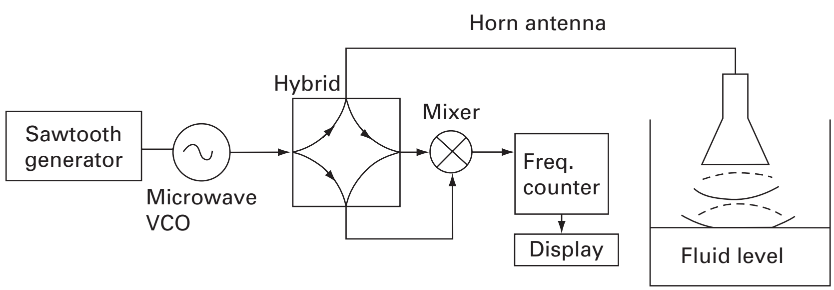

Figure 21.3 Radar fluid level detector.

The frequency of the tone at the mixer output is given by the frequency difference of the two signals arriving at the mixer inputs. One of these signals is delayed by the round-trip time to the fluid surface, so the lower the level in the tank, the higher the output frequency of the mixer. Note that this radar cannot determine velocity; if the fluid level were changing rapidly, it would create a

Doppler shift that could not be distinguished from a change in fluid level. (Range and velocity are ambiguous for this radar.)

21.1.3 Pulse-Doppler radar

The radar of Figure 21.1 can be modified to sense velocity as well as position. The resulting *pulse-Doppler radar*, when mounted on an aircraft, can distinguish moving targets on the ground from the much stronger "ground clutter" produced by stationary objects. The earliest form of pulse-Doppler radar, called "MTI", for *moving target indicator*, used a mercury acoustic delay line as analog storage for the train of echos received after each pulse. The stored signal was subtracted from the echoes produced by the next pulse to remove the echoes from fixed targets. (This type of delay line, when configured as a recirculating shift register, was used around 1950 as the memory in some of the earliest digital computers.)

Pulse-Doppler radar really became practical with the advent of frequency-stable microwave signal sources (klystron amplifiers, rather than high-power magnetron oscillators), digital storage capability, and digital signal processing. These advances allow coherent signal averaging for increased sensitivity, as well as actual determination of target velocity. Successive received echoes, after being mixed to a low IF frequency, are stored, preserving both amplitude and phase, and then the stored echos are analyzed to see how the phase changes from one pulse to the next. A moving target causes a progressive phase change (Doppler shift) from which the line-of-sight velocity can be determined. When used as a search radar, successive received signals are stored separately for every range. Fourier analysis of the sequence of sampled IF voltages for each range determines whether there is a target at that range and, if so, determines its velocity. A problem that often arises with pulse-Doppler systems involves a range-velocity ambiguity. If target velocities are in the range $\pm v_{\max}$, the corresponding Doppler shift range is $\pm 2f\, v_{\max}/c$, where f is the radar frequency. This produces a total received bandwidth of $4fv_{\max}/c$. If we are to unambiguously determine the frequencies present in this band, the sampling rate, which is the pulse rate, must be at least $8f\, v_{\max}/c$. A high radar frequency and a high maximum velocity may dictate a pulse rate that restricts the maximum unambiguous range. (The echo from a target beyond the maximum unambiguous range arrives after the next pulse has been transmitted, as if it had been reflected from a close-in target.) This problem is discussed in Chapter 26 in the context of planetary radar astronomy.

21.2 Radar classification

Many radar systems are quite different from the classic configuration of Figure 21.1. Radar systems are conveniently classified according to several key system characteristics:

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(a) Monostatic or bistatic:
A monostatic radar uses a single antenna for both transmitting and receiving, while a bistatic radar uses two separate antennas. An obvious advantage of the bistatic configuration is that it requires no TR (transmit/receive) switch. A military advantage is that only the transmitting site, which can be unmanned, is vulnerable to attack by a radiation-seeking missile. Bistatic and multistatic radar systems can also be passive, using "transmitters of opportunity," such as television broadcast transmitters.

(b) Coherent (Doppler) or incoherent:
Radars with stable frequency sources for the transmitter and the receiver L.O. can use coherent integration to extract a signal from noise. They can also use the Doppler effect to measure the line-of-sight velocity of a moving target. Incoherent radars detect the power of each pulse, making signal averaging less efficient. They cannot directly detect target velocity though they can, of course, infer velocity by observing successive changes in range.

(c) Pulse or cw:
Short pulses can distinguish closely separated targets, providing range resolution. Long pulses, whose limit is cw (continuous wave), can distinguish target velocity using the Doppler shift. Nevertheless, coherent trains of short pulses and phase-modulated long pulses can provide both range and Doppler resolution.

(d) Search or track:
Aircraft traffic monitoring is normally done with a monostatic search radar, using a rotating antenna that produces a beam pattern that is tall in elevation but narrow in azimuth. Echos from aircraft produce "blips" on a PPI (plan-position indicator) screen which is refreshed on every rotation of the antenna. A tracking radar uses servo motors to keep the antenna pointed at a single, usually fast-moving target.

(e) Mechanical or electronic scanning:
A *phased array* uses a closely spaced two-dimensional array of antenna elements, usually dipoles. Each dipole is equipped with a programmable phase shifter, allowing the array to form a beam in an arbitrary direction. If the signals received from the individual dipoles are made available to the signal processor, the processor can simultaneously form several independent receiver beams. This allows tracking of multiple targets or searching of multiple zones. Phased arrays with three or more faces of elements can be completely stationary.

(f) Detection or imaging:
Conventional radars simply detect the position and/or velocity of targets. Imaging radars make photographic-like reflectivity maps. Imaging systems include airborne side-looking (*synthetic aperture*) radars that map a strip of land parallel to the flight path and "planetary radar" (delay-Doppler) systems that image planetary objects and low-orbit artificial satellites.

21.3 Target characteristics and echo strengths

21.3.1 Radar cross-section

The strength of an echo depends on the nature of the target and on its range from the radar. In general, metal targets with dimensions commensurate with the wave length are good reflectors, especially when their geometry includes linear features. For example, airplane propellers act as dipole re-radiators. A flat metal plate produces a strong specular reflection, if perpendicular to the direction of the radar, but edge currents also produce other, less directional, echoes. Radar engineers define the *radar cross-section* (RCS) of a target, σ, as the collecting area of an object that would produce the target's observed echo strength while *isotropically* re-radiating the intercepted incident radiation. Radar cross-sections for practical targets, such as aircraft, depend on the aspect angle of the target as well as on the target's usually complicated geometry. RCS calculations for such targets are carried out using finite-element electromagnetics modeling programs. Exact expressions have long since been derived for the radar cross-sections of simple objects. A sphere, since it has no aspect angle dependence, makes a good calibration target for measuring radar sensitivity. The RCS of a metal sphere of radius a is just πa^2 (its geometric cross-sectional area) when a is greater than about $5\lambda/\pi$. But when a is less than about $\lambda/(4\pi)$, the RCS is given by the formula: $\sigma = \pi a^2\ [9\ (2\pi a/\lambda)^4]$. Note that the cross-section is proportional to a^6 and to λ^{-4} in this small-target or *Rayleigh scattering* regime.[2]

21.3.2 The radar equation

The definition of radar cross-section lets us write a simple expression for the flux density of a backscattered echo as it returns to the radar antenna

$$S_{\text{received}} = S_{\text{incident}}\sigma/(4\pi R^2), \tag{21.1}$$

where R is the distance to the target. This equation just says that the supposed isotropically scattered power, which is given by $S_{\text{incident}}\ \sigma$, produces a flux density determined by the area of a target-surrounding sphere whose radius is the distance to the radar. From antenna considerations discussed in Chapter 20, we can write the incident flux in terms of the transmitted power as $P_{\text{trans}}\ G/(4\pi R^2)$, where G is the gain of the antenna. Of course, this assumes the antenna is pointed directly at the target. Likewise, we can express the capture area of the antenna by $A_{\text{eff}} = \lambda^2 G/(4\pi)$. Putting this together, we obtain an expression for the received power, P_{rec}, which is just $S_{\text{incident}}\ A_{\text{eff}}$ or

[2] The reason the sky is not black is that the molecules that make up the atmosphere scatter direct sunlight to produce diffuse sky light. The λ^{-4} dependence of Rayleigh scattering causes sky light to be blue and sunsets to be red.

$$P_{\text{rec}} = [P_{\text{trans}}G/(4\pi R^2)][\sigma/(4\pi R^2)][\lambda^2 G/(4\pi)] = P_{\text{trans}}G^2\sigma\lambda^2/[(4\pi)^3 R^4)]. \quad (21.2)$$

This relation, however written, is known as the *radar equation*, and shows the inverse fourth-power dependence on range. For an aperture antenna, e.g., a dish antenna, the effective aperture is given by $A_{\text{eff}} = \eta_{\text{ap}}\, \pi(D/2)^2$ where D is the dish diameter and η_a, the aperture efficiency, is usually around 0.5. Using this, we can write the radar equation as

$$P_{\text{rec}} = P_{\text{trans}} \frac{\pi}{64} {\eta_a}^2 \left(\frac{D}{R}\right)^4 \frac{\sigma}{\lambda^2}. \quad (21.3)$$

21.3.3 Distributed targets

Targets such as aircraft subtend a solid angle much smaller than the radar beam, but distributed targets, such as rain in the atmosphere or free electrons in the ionosphere, can be much larger than the beam. In these cases the beam actually defines the extent of the target, in as much as there will be radar echoes from throughout the entire volume of the beam. When the individual scattering objects (raindrops and electrons in the above examples) are randomly distributed in space (*incoherent scattering*), the power received by the radar is the sum of the powers of the individual scatterers.[3] For this kind of target, the radar equation must be modified. We first define σ_V, the radar cross-section per unit volume of the target, to be $\sigma_V = n\sigma$, where n is the volume density of scattering particles and σ is the radar cross-section of an individual particle. For raindrops, σ is about 41% of the Rayleigh scattering cross-section given above for metal spheres. For individual electrons, $\sigma = [e^2/(4\pi\epsilon_0\, m_e c^2)]$, where e and m are the charge and mass of an electron and c is the speed of light. If we extend the arguments we used for a single particle centered in the beam to include each volume element throughout the beam, we find that

$$P_{\text{rec}} = \frac{P_{\text{trans}}\lambda^2\sigma_V c\tau}{(4\pi)^2 R^2} \frac{1}{4\pi} \int G^2(\phi,\theta)\,\mathrm{d}\Omega. \quad (21.4)$$

In this radar equation τ is the pulse length, so $c\tau$ is the range depth of the target volume. The cross-range area of the beam is taken into account by integrating over solid angle times R^2, the square of the range. Note that, because the beam determines the target size, the return echo of a distributed target is proportional to R^{-2}, rather than R^{-4}. The last term in Equation (21.4), $(4\pi)^{-1} \int G^2\, \mathrm{d}\Omega$, is the mean square gain, also called the *backscatter gain* of the antenna.

[3] The electric field of the radar echo is the sum of the field contributions from the N individual scattering particles illuminated by the beam. If the scatterers have random positions, their E-field contributions have random phases and add in the fashion of a random walk, causing the total E-field echo to be proportional to $\sqrt{N}$. The received power, proportional to the square of the received E field, is therefore proportional to N.

21.4 Pulse compression

Short pulses produce good range resolution, but shortening the pulses reduces the sensitivity of the radar unless the peak power (pulse power) can be increased to maintain the same average power. However, the peak power of any transmitter is eventually limited by voltage breakdown or other device limitations. This constraint led to the development of *pulse compression* schemes in which the transmitted pulses are modulated in such a way that the echoes can be "bunched up" by the receiver as if they had started out as short pulses. The received echo is passed through an appropriate matched filter (often a digital processor) whose output, for a point target, will be a narrow pulse.[4]

One method for pulse compression uses phase coding. The transmitted pulse is divided into N contiguous equal intervals. Each interval or *baud* is assigned a phase value of zero or 180 degrees, which will be the relative phase (and therefore polarity) of the transmitted signal during that interval. These assignments specify the code. Normally every transmitted pulse will have the same code. Figure 21.4 shows how such a code is used for pulse compression. At the receiver, the IF signal from a point target will be a replica of the transmitted pulse. This IF signal travels down a delay line. Taps along the delay line have a delay spacing equal to the baud length. Signals extracted at these taps are weighted (multiplied) by coefficients with values $\pm$ 1, whose order duplicates the code. The weighted signals are added together to form the pulse compressor output.

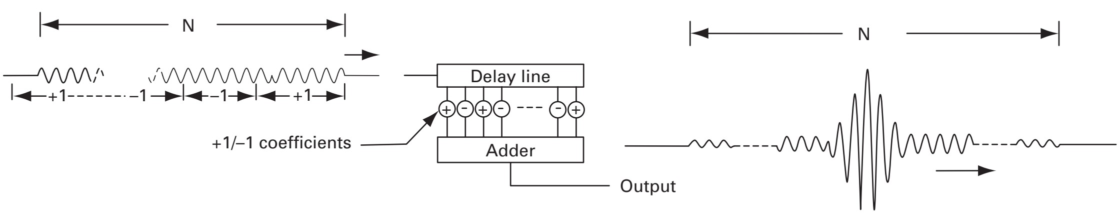

Figure 21.4. A compressor for phase-coded pulses.

Note that, for one position of the signal in the delay line, the multipliers will exactly "undo" the coding, giving the same phase to each of the N input signals to the adder. This alignment will peak at the middle of one baud, as the signal propagates down the delay line, attaining a peak amplitude N times greater than the amplitude of the uncompressed input signal. When the signal in the delay line is not in alignment with the code, there will be some output from the adder unless the code is such that the addends sum to exactly zero. There is no code with this property, though the *13-baud Barker code*, $+ + + + + - - + + - + - +$, produces

[4] In general, high range resolution requires that the transmitted signal have a large bandwidth. This bandwidth can be produced by transmitting either short pulses or appropriately modulated long pulses.

("sidelobe") values of only +1 or 0 when the code is not aligned and +13 when it is aligned. When longer codes are needed, it is common to use pseudorandom codes. In this case, the sidelobe levels are determined by random walk statistics; the close-in sidelobes will be less than the main lobe by a factor on the order of $\sqrt{N}$.

An analog surface acoustic wave (SAW) processor can do all the operations shown in Figure 21.4, but it is now more common to use digital processing. Typically, baseband signals are produced by mixing the IF signal, centered at ω_{IF} with $\cos(\omega_{\mathrm{IF}}\ t)$ and $\sin(\omega_{\mathrm{IF}}\ t)$. The resulting I and Q signals are each furnished with a tapped digital delay line/adder pulse compressor. The outputs from the two compressors can be squared and added together for immediate detection, or they can first be Doppler processed.

Pulse compression has other benefits. Because the rapidly changing code spreads the signal's spectrum, it lowers the spectral density. This gives a military radar signal an element of stealth; being less noticeable, it is less likely to provoke a hostile reaction, such as an antiradiation missile. By using different codes with low cross-correlation properties, multiple radars can share the same frequency band. This code-division multiple access property is also used in CDMA cellular telephone systems and in the GPS satellite system, where multiple users (the callers or satellites) share the same band. The GPS system also makes use of the pulse compression property to achieve the high time resolution needed for accurate range resolution (see Chapter 25).

21.5 Synthetic aperture radar

As described above, a phased-array antenna combines the signal from many individual elements to form a beam pattern. Consider a phased array directed toward a point source of cw radiation whose frequency is precisely known. The analog voltage from each element is given a programmed analog phase shift. These voltages then combine to form a vector sum. Note, however, that it is also possible to digitize the voltage (phase and amplitude) from each element and then *compute* the vector sum. If we do this, no hardware phase shifters are needed since, before summing, we can multiply each voltage, V_i by a phase factor, $e^{j\phi_i}$, to point the beam in any desired direction. This can be repeated with other sets of phase factors to form a host of beams from the one set of digitized voltages. The echo powers in these beams then form an image if there are multiple targets (assuming they all radiate at the same frequency). Finally, note that we could do all of this with only a *single* array element by moving the element successively from one position to another, recording the voltage (amplitude and phase) at each position. If the set of positions is identical to the element positions in the original phased array, we have effectively used a single small element to *synthesize* a steerable large-aperture antenna.

You can see that this all works out in a radar situation, where all the targets are illuminated by a coherent transmitter signal. The transmitter also provides a

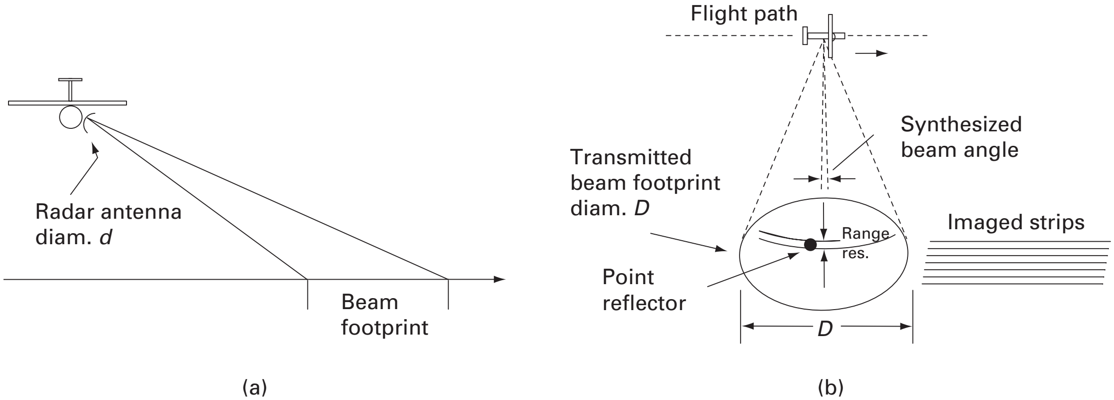

Figure 21.5. Side-looking synthetic aperture radar: (a) elevation view; (b) plan view.

phase reference for the receiver so it is practical to measure both the amplitude and the phase of the return signal at the antenna element. This is the general idea of *synthetic aperture radar* (SAR).[5] Here we assumed that the target was stationary and that the element was stationary, except while being relocated after each voltage measurement. In practice, most SAR techniques involve continuous motion of the elemental antenna, relative to the target. This continuous motion makes it possible to explain the synthesis procedure in terms of Doppler shifts. In addition, most SAR systems transmit short pulses to provide resolution in the down-range (pointing) direction, and use one-dimensional aperture synthesis to provide resolution in the cross-range direction. One of the most common SAR systems is the side-looking radar shown in Figure 21.5.

A small radar antenna of diameter d points outward and downward from the aircraft, which flies a straight-line course. The illuminated patch on the ground, called the *radar footprint*, has a large diameter, D. Since the beamwidth from the radar antenna is given by λ/d radians, we see that D is given by $D=R\lambda/d$, where R is the distance of the footprint from the aircraft. Range resolution is obtained by simply using short radar pulses. In the cross-range direction, the system synthesizes an antenna whose extent in the direction parallel to the flight path is D, since that is the dimension of the footprint that passes over any target feature. The angular size of the resulting synthesized beam is given by $\delta\theta=\lambda/D$. Substituting $D=R\lambda/d$ produces $\delta\theta=d/R$, and the cross-range resolution is therefore $R\delta\theta=d$, the diameter of the small radar antenna. It is interesting to note that the resolution is independent of λ and that making d smaller improves rather than degrades the resolution. Many range strips are observed simultaneously. You can find fascinating side-looking radar (and side-looking sonar) images by searching the internet. A "delay-Doppler" SAR system to image

[5] Synthetic aperture mapping is also possible when the target is not illuminated by coherent radiation, but emits its own wideband radio "noise." In this case, aperture synthesis requires at least two antenna elements. This interferometry technique, used by astronomers to make maps of the radio sky, is discussed in Chapter 26.

planetary surfaces is discussed in Chapter 26. Another interesting but difficult technique is *inverse synthetic aperture radar* (ISAR) where the aspect of the target (which might be a ship bobbing on the ocean) is variable, not known a priori, and must be inferred from the radar returns as part of the data processing.

21.6 TR switches

Monostatic radars, which transmit and receive with the same antenna, require a TR (transmit-receive) switch. In most radar applications the desired echo arrives so soon after the pulse is transmitted that the TR switch (also known as a *duplexer*) must be electronic rather than a mechanical. Here we will first look at self-duplexing radar techniques based on the use of circular polarization or circulators, then at standard TR switch circuits, and finally at RF electronic switches in general.

21.6.1 Self-duplexing radar techniques

If a radar transmits a circularly polarized signal, reflection by the target changes the sense of polarization from left-hand to right-hand or vice versa. Circular polarization can be produced by transmitting simultaneous crossed linear polarizations 90° out of phase. Figure 21.6 shows how a 90° hybrid not only produces circular polarization of one sense but also routes received circular polarization of the other sense (the return signal) into the receiver.

Note that the x-dipole and the y-dipole, together with the hybrid, are really just equivalent to two separate antennas having opposite circular polarizations. In practice, the isolation between the transmitter and receiver in this scheme is usually no better than about 30 or 40 dB, so a limiter or SPST electronic switch (a *monoplexer*) is installed at the receiver input to protect it from burnout. A waveguide version of this circuit uses a *turnstile junction*, the microwave component shown in Figure 21.7. It is classified as a six-port junction because the round waveguide supports two independent modes: x and y or RCP and LCP or any other pair of orthogonal elliptical polarizations. When two opposite

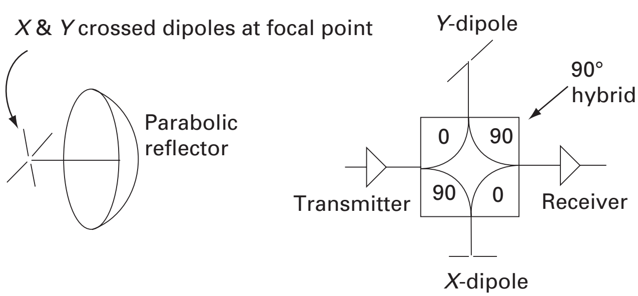

Figure 21.6. A self-duplexing radar using circular polarization.

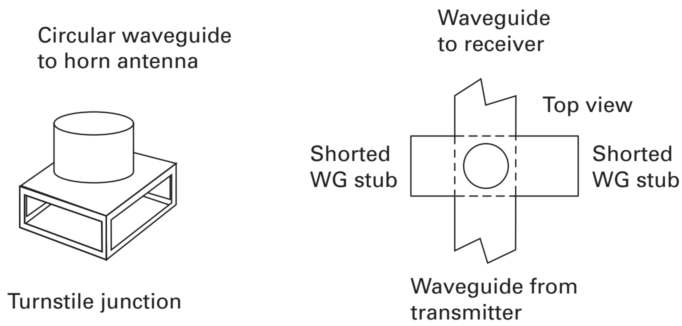

Figure 21.7. A waveguide turnstile junction combines the functions of the hybrid and crossed dipoles.

rectangular ports are fitted with shorts of appropriate lengths, the resulting four-port network is equivalent to the pair of dipoles and hybrid of Figure 21.6. The turnstile junction is described in Volumes 8 and 9 of the Rad. Lab. Series. The transmitter and receiver are connected to the remaining rectangular ports (the pair without shorts) while the antenna, usually a feed horn, is connected to the round waveguide.

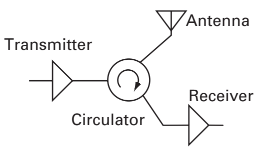

Figure 21.8. A circulator used as a TR switch.

A true self-duplexing circuit, shown in Figure 21.8, uses a circulator. The circulator has the property that a signal injected at the transmitter port will exit via the antenna port while a signal injected at the antenna port will exit through the receiver port. (If a signal were injected at the receiver port, it would exit through the transmitter port.) This nonreciprocal action depends on transmission through a nonreciprocal medium which, for the circulator, is a ferrite material, biased by the field of a permanent magnet. This elegant TR system is limited by available circulators to powers of tens of kilowatts.

21.6.2 TR switching devices and circuits

The classic TR circuits developed during WWII use gas discharge tubes or spark gaps and are self-activated by high-power RF on the transmission line. Lower power versions use PIN diode switches, turned on by an external bias circuit. (The radar has a timing generator providing pulses that (1) turn on the TR switch for transmitting, (2) pulse the transmitter, and (3) turn off the TR switch for receiving.) Gas discharge tube switches are usually built into a short piece of waveguide and come in two types: TR and ATR (anti-TR). The distinction is as follows: on transmit a TR device ionizes and presents a low-impedance shunt across the line. The ATR device also ionizes on transmit but it presents a low impedance in series with the line. (Reference [3], Volume 14 of the Rad. Lab. Series, is devoted mostly to these tubes.) There are two general classes of TR circuits, branch line TR switches and balanced TR switches. The former uses segments of transmission line while the latter uses hybrids.

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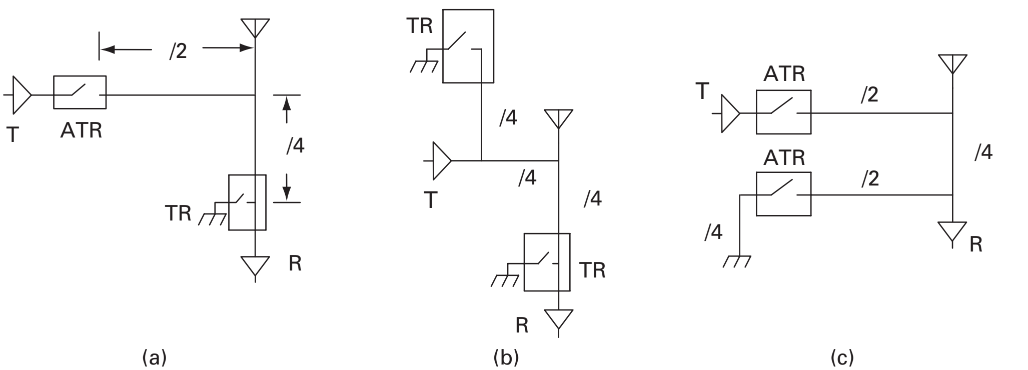
21.6.2.1 Branch line TR switches

Figure 21.9 shows some standard branch line TR Switch circuits using TR, ATR, or both TR and ATR tubes (or PIN diode equivalents).

In the circuit of Figure 21.9(a), the switches are shown in the nonconducting (transmitter off) position. The open ATR switch is connected to the antenna by a half-wave line so it presents the same open circuit to the antenna-to-receiver line. Likewise, the open TR switch does not disturb the antenna-to-receiver line. On transmit, the TR switch places a protective short circuit at the receiver input. This short circuit is transformed by the quarter-wave line into an open circuit which does not affect the connection between the transmitter and the antenna. At high frequencies the switches contain some nonzero path lengths which form part of the half-wave or quarter-wave lines. For low-frequency designs the half-wave lines can be reduced to zero length and the quarter-wave lines can be replaced by lumped-element impedance inverters. You can see from the ways that the TR and ATR elements are used that their names are somewhat arbitrary.

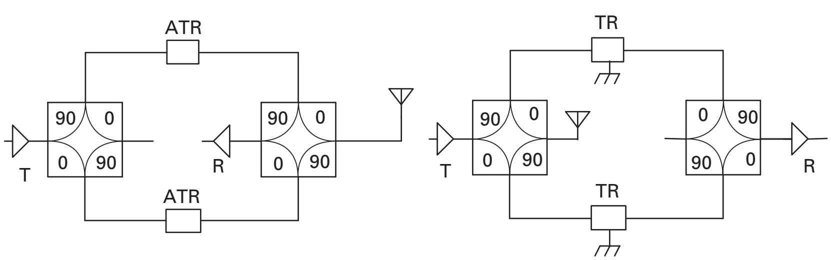
Figure 21.9. Branch line TR switches.



21.6.2.2 Balanced duplexers

Balanced duplexers use hybrids and can have wider bandwidths than the branch line circuits shown above (though more elaborate branch line circuits can have wider bandwidths). Both balanced duplexer circuits shown in Figure 21.10 use two 90° hybrids.

Figure 21.10. Balanced TR switches.



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21.7 Diode switches

A single diode shunt switch circuit is shown in Figure 21.11.

Positive control voltage gives the diode a forward bias current to produce a low dynamic resistance, dV/dI. Negative control voltage turns the diode off, making its dynamic resistance very high. Let these dynamic resistance values be denoted respectively by r and R. You can verify (Problem 21.5) that the transmission values (power out/power available) for the switch of Figure 21.11 are given by:

$$\text{Isolation (forward biased state)} = 4r^2/Z_0^2$$
$$\text{Transmission (reverse biased state)} = 1/(1 + Z_0/R).$$

Note from these expressions that you could favor better isolation or lower insertion loss by transforming the line to have a larger or smaller Z_0 at the diode location. Better performance can be obtained with ladder networks analogous to multisection filters. Figure 21.12 shows how a shunt switch can be combined with a series switch to form a two-element ladder network. Isolation is improved since any signal leakage across the open series switch is shorted to ground by the closed shunt switch.

It is often more convenient to use shunt diodes than series diodes, since shunt diodes are easier to bias and to heat sink. Impedance inverters can transform series elements into shunt elements, as we saw when designing coupled

Figure 21.11. Shunt diode switch.

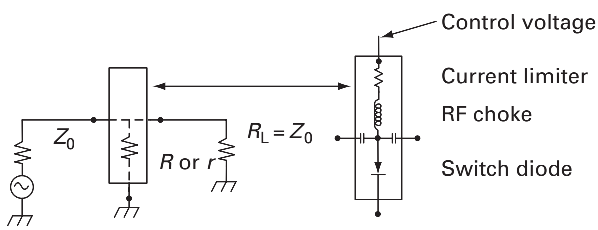
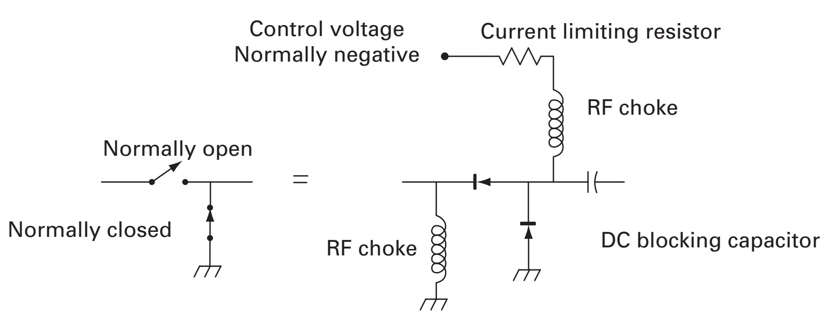


Figure 21.12. Series-shunt diode switch.



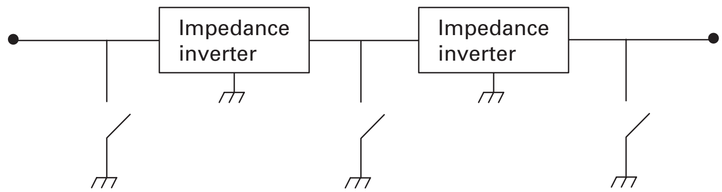


Figure 21.13. Three-section switch using only shunt sections.

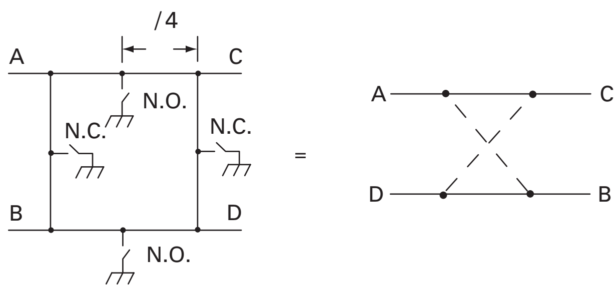


Figure 21.14. Transfer switch using quarter-wave transmission lines.

resonator filters. The switch circuit of Figure 21.13 uses two impedance inverters and three shunt switches.

The switches discussed above are all SPST switches. More complicated switches can be built up from the elementary SPST switch, but special designs can often be made such as the four-diode transfer switch shown in Figure 21.14. When the switches are in the indicated normal positions, the shorted quarter-wave lines appear as infinite-impedance shunts and do not disturb the transmission paths from A to C and from B to D. When the switches are reversed, transmission is from A to B and from C to *D*.

21.7.1 Diodes for RF switching

When ordinary diodes are used in these switching applications, the biases must be enough to keep the diode in the desired state. In particular, when the diode is off, the reverse bias voltage must be greater than the peak RF voltage and, when the diode is on, the forward bias current must be greater than the peak RF current. However, the *PIN diode*, a sandwich of p-type, intrinsic, and n-type semiconductor material, has the remarkable property that, for RF switching, the bias current and bias voltage values can be less than the corresponding RF current and voltage by perhaps an order of magnitude. The operation of the PIN diode (see reference 5) depends on having a large (small) stored charge in the intrinsic region when the diode is on (off). At high frequencies, the time between electric field reversals is much less than the transit time through the intrinsic region, determined by diffusion and drift, so the diode remains in the on (off) state. Finally, a word of caution: diodes, since they are nonlinear circuit elements, have the potential to distort a signal. In particular, they can create intermodulation products between the various signals in a complicated spectrum. For critical applications, e.g., receiver band-changing, diode switches must be turned off and on hard enough to keep any generated intermodulation products at a negligible level.

21.8 Radar pulse modulators

Any cw transmitter can generally be used as a pulse transmitter if a *pulse modulator* is added to provide the rapid turn on and turn off. Tube-type amplifiers can be operated with much higher instantaneous powers when they are pulsed. Tubes are primarily limited by their maximum anode dissipation (heat removal); the dissipation can be the result of either modest cw operation or high-power pulse operation. A cw amplifier can be converted for pulse operation by changing the output matching circuit in order to present a lower load resistance to the tube. Some tubes are available in special pulse-rated versions; they are fitted with high-emission cathodes. Gridded tubes (triodes, tetrodes, and pentodes) can be pulsed by switching the grid bias from negative, for pulse-off, to positive, for pulse-on. The negative bias keeps the tube completely turned off between pulses. Since the grid voltage and current are much smaller than the plate voltage and current, grid control requires only low-power circuitry compared to anode control. At microwave frequencies, magnetrons and klystrons replace gridded tubes. Magnetrons have no control element and therefore require high-power anode pulsers. Klystrons may or may not have a modulating anode ("mod anode") by which the beam current can be cut off. If not, they need high-power pulsers.[6]

Transistor amplifiers, unlike tube amplifiers, can make very little trade-off between duty cycle and peak power. Transistors suffer one type of breakdown or another when operated beyond their maximum continuous ratings. A high-power transistor amplifier for pulse service might differ from a cw amplifier only in that it will dissipate less heat (from the reduced duty cycle) and can therefore get by with a smaller heat sink.

No matter how an amplifier is pulsed, the power supply must furnish high-power pulses with minimum voltage droop. Duty cycles of pulsed transmitters are usually much less than unity so, in addition to at least one switching element, pulse modulators (pulsers) contain some form of energy storage element(s). The simple pulser circuit shown in Figure 21.15(a) stores energy in a capacitor.

[6] An air traffic control radar might have a peak power output of 2 MW and an efficiency of 50%. A klystron tube in this service could require 50 kV pulses at 80 amperes.

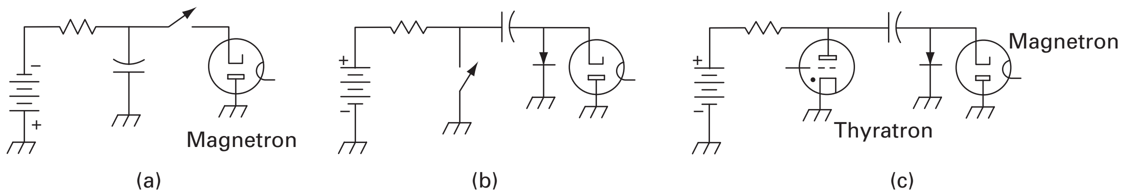


Figure 21.15. Capacitor discharge pulser.

In this circuit the tube (magnetron, klystron, or whatever) is shown as requiring negative voltage. Microwave tubes often use negative supply voltage applied to their cathodes because it is convenient to ground the external heat-dissipating anode. The version of the circuit in (b) allows one side of the switch to be grounded, which is another convenience. The diode provides a charging path for the energy storage capacitor. The circuit of Figure 21.15(c) uses a thyratron (vacuum tube version of the SCR) as the switch.

The simple capacitor discharge modulators in Figure 21.15 have several disadvantages: The voltage droops during the pulse. The droop can be reduced by increasing the size (weight, and cost) of the capacitor. Not much of the stored energy is used. Even if 10% voltage droop is permitted, only 20% of the stored energy is used for each pulse. This might be compared to a car which would not run well if the fuel tank was less than 80% full. They are expensive and heavy, a particular disadvantage for airborne equipment. Despite these disadvantages, capacitor banks are often used, as in the 430-MHZ pulse transmitter used for ionospheric research at the Arecibo Observatory, because a more efficient circuit, the line modulator discussed below, does not easily provide the flexibility needed to change the pulse width.

21.8.1 Line modulators

A length of transmission line (with the far end open) has capacitance and can therefore store electrostatic energy. When the line is discharged into a resistive load equal to its characteristic impedance, it will supply a perfect rectangular pulse rather than a drooping exponential pulse. The constant pulse amplitude during discharge is maintained by the distributed inductance of the line acting together with the distributed capacitance. In Figure 21.16(a), the line is a piece of coaxial cable, replacing the energy storage capacitor. As before, the tube is supplied with a negative pulse. A diode provides a path to recharge the line. Often the load has a higher impedance than the characteristic impedance of the line, and a pulse transformer is required.

The line supplies a pulse at half the charging voltage since, during the pulse, the charging voltage evenly divides between the load and the equivalent source resistance. The duration of the pulse is the time taken for the current to make a round trip through the line. At the end of the pulse the line is totally discharged; all the stored energy is delivered on every pulse. Figure 21.16(b) shows successive plots of the line's voltage and current distributions. In order to store more

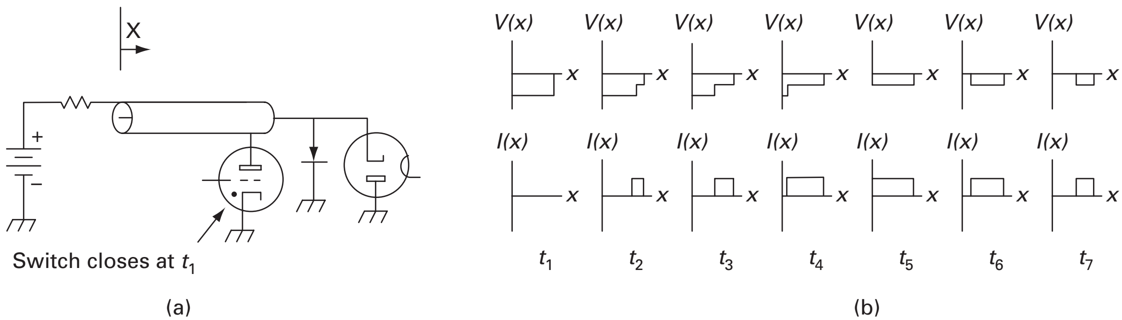
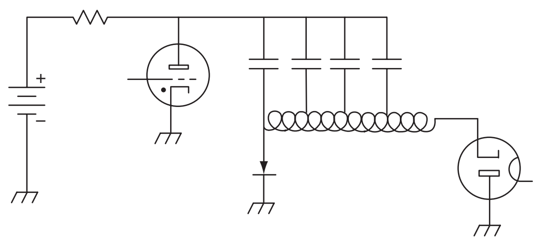


Figure 21.16. (a) Line-type modulator; (b) line voltage and currents.

Figure 21.17. Pulser using an artificial transmission line (pulse-forming network).



energy, it is common to use an "artificial transmission line" or pulse forming network (PFN) which is a ladder network of inductances and capacitances. A four-section network is shown in the modulator circuit of Figure 21.17.

The network looks like a lowpass filter and it is. Its cutoff frequency is given by $\omega^2 = 4/[LC]$. For frequencies well below cutoff, the network behaves like a transmission line with $Z_0 = \sqrt{L/C}$. Here L and C are in henries and farads rather than henries/meter and farads/meter as in the distributed element transmission line. The one-way time delay on this lumped line is $\sqrt{LC}$ seconds/section.

Let us consider a numerical example: Let us use four sections as in the figure above. Suppose we need a 1 microsecond pulse at 10 kV and 10 amperes. The voltage and current require that $Z_0 = \sqrt{L/C} = 1000$ ohms. We will use four sections which, for the desired one-microsecond round-trip delay, requires that $8\sqrt{LC}$ be equal to 10^{-6}. These impedance and time delay equations are satisfied by $L = 125$ μH and $C = 125$ pF. We can verify that the energy stored in the line is indeed equal to the energy delivered by the pulse. The latter is just $(IV)\tau = 10 \times 10\,000 \times 10^{-6} = 0.1$ joule. The former, remembering that we must charge the line to 20 000 V, is $CV^2/2 = 4(125 \times 10^{-12}) \times 20\,000^2/2$ which is also 0.1 joule. As often happens in filter design, these are not particularly practical values; real inductors of 125 μH may well have distributed capacitances that are not negligible compared with 125 pF. We can build the line for a lower

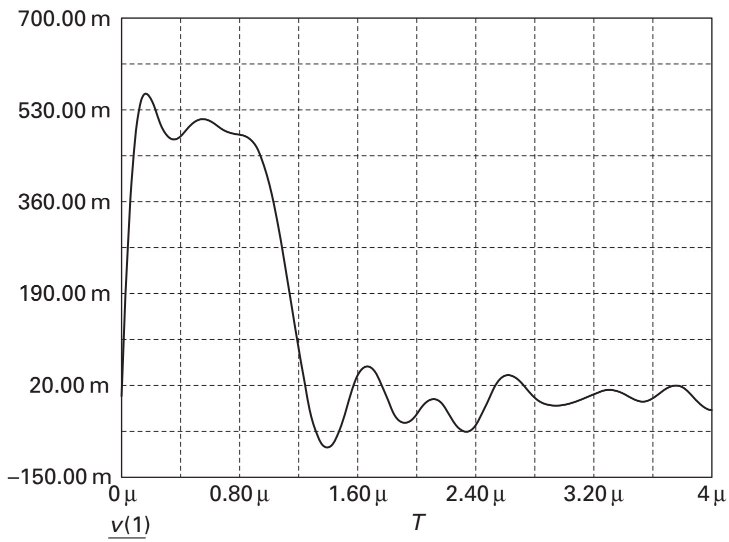


Figure 21.18. Waveform produced by a four-section pulse forming network.

impedance and use a pulse transformer between the line and the magnetron. If we lower the line impedance to 100 ohms, the L and C values become 12.5 µH and 1250 pF, values that are more practical. Using these values, a SPICE simulation of the discharge produced the voltage waveform shown in Figure 21.18. The voltage scale is normalized, i.e., the capacitors were charged to one volt so the nominal pulse voltage is 0.5 volts. Most pulse forming networks can be tuned slightly to improve the pulse shape; the artificial transmission line is, after all, only an approximation to an ideal transmission line.

The line modulator uses all the stored energy on each pulse but, precisely because of this virtue, deserves a more sophisticated charging circuit than the resistor shown in the circuits above. Remember that when a capacitor is charged through any resistive path from empty (no energy) to $CV^2/2$, the resistor will dissipate this same amount of energy, $CV^2/2$. Here the charging resistor, no matter what value, would dissipate half the power consumed by the radar. The solution to this problem is to charge the line through an inductor instead of a resistor. Figure 21.19(a) shows the voltage waveform on a capacitor as it is *resonantly charged* through an inductor.

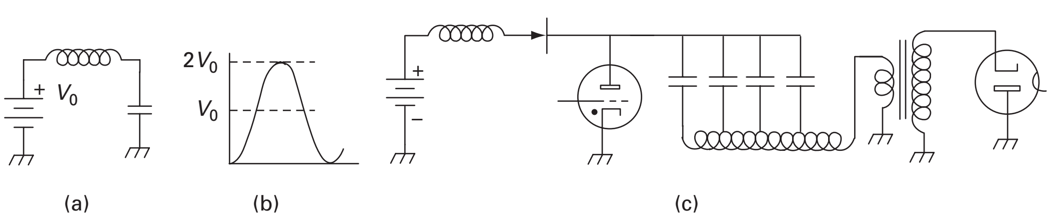


Figure 21.19. Resonant charging.

The voltage is a sinusoid, building up to a maximum of twice the supply voltage. The modulator can be triggered just as the voltage reaches this maximum. The brief pulse discharges the line and the charging curve begins anew. It would seem that the PRF is therefore determined rigidly by the charging time but, if a diode is put in series with the inductor, the charging stops at the maximum voltage and the next pulse can occur anytime. The resonantly charged modulator, with the diode and a pulse transformer, is shown in Figure 21.19(b). Note that the primary of the pulse transformer provides a charging path, eliminating the diode originally in parallel with the magnetron. Also remember that, because of the resonant charging, the supply voltage needs only to be half of the line charging voltage. Line modulators present less risk to tubes than partial-discharge capacitor modulators because there is less stored energy available when an arc occurs in the tube.

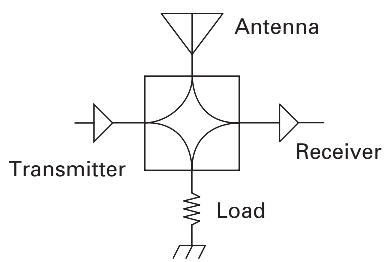
Problems

Problem 21.1. Find the radar cross-section of a flat metal plate of area A that is exactly perpendicular to the radar beam. Assume that $A \gg \lambda^2$. Hint: treat the plate as an aperture antenna with gain $4\pi A/\lambda^2$.

Problem 21.2. Find the lobe pattern produced when a point target is observed by a pulse compression radar using the 13-bit Barker code of Section 21.4. (Convolve the code with itself after padding both ends with zeros.)

Problem 21.3. Why is the side-looking radar antenna positioned to look to one side of the aircraft rather than straight down?

Problem 21.4. If we try to use the hybrid of the circular polarization duplexer as a circulator, we might consider the TR circuit shown below.



Assume the antenna, transmitter, receiver, load, and hybrid all have the same characteristic impedance. This circuit at least protects the receiver from the transmitter. What are its disadvantages (a) when transmitting, and (b) when receiving?

Problem 21.5. Verify the expressions below for $P_{out}/P_{available}$ for the circuit of Figure 21.11.

Diode state	Switch state	$P_{out}/P_{available}$
Foward biased	Isolation	$4r^2/Z_0^2$
Reversed biased	Transmission	$1/(1+Z_0/R)$

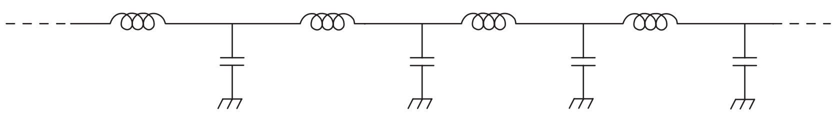
Problem 21.6. Apply your circuit analysis program (Problem 1-3) to the transfer switch circuit of Figure 21.4. Assume a 50-ohm load is connected to Port C, a 50-ohm generator is connected to Port A, and the transmission line sections have a 50-ohm characteristic impedance. Assume also that the internal switches are ideal. Find the transmission coefficient (in dB) over the frequency range from half the design frequency to twice the design frequency. Hint: the closed switches divide the circuit in two so you can ignore the bottom half.

Problem 21.7. Explain the operation of the balanced duplexers shown in Figure 21.10. What restrictions, if any, are there on the lengths of the interconnecting transmission lines?

Problem 21.8. (a) Show that when an uncharged capacitor is brought to potential V by connecting it through a resistor to a voltage source V, the energy supplied by the source is twice the energy deposited in the capacitor (CV^2 rather than $CV^2/2$).

(b) The charging efficiency in (a) is only 50%. Find the efficiency when the capacitor initially has a partial charge, i.e., when the capacitor is initially charged to a voltage αV, where $\alpha < 1$.

Problem 21.9. (a) Find the characteristic impedance of the artificial transmission line shown below. This impedance, Z_0 (which is complex), can be found by adding another LC section to the properly terminated line and noting that the new impedance must still be Z_0.



(b) Use the expression for Z_0 to show that the line has a cutoff frequency, $\omega_c = 2/\sqrt{LC}$, above which signals are reflected rather than transmitted.

Problem 21.10. Show that when $\omega \ll \omega_c$, the propagation delay per section for the artificial transmission line of Problem 21.9 is given by $\tau = \sqrt{LC}$.

Problem 21.11. (a) In a *conial scan* tracking radar, the antenna's feed horn is tilted slightly off center and mechanically rotates around the axis of the parabolic reflector. The antenna beam therefore executes a continuous tight conical scan, centered on the target. Many pulses are transmitted during the course of each scan. Draw a block diagram of circuitry to furnish x and y cross-range position error signals to the antenna's drive system in order to keep the antenna centered on the target. (b) Consider a *monopulse* tracking radar in which the antenna is effectively five antennas: one pointed directly at the target, one slightly above, one slightly below, one slightly to the left, and one slightly to the right. Draw a block diagram of circuitry to combine these five signals to provide x and y error signals.

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