J. Webster (ed.), *Wiley Encyclopedia of Electrical and Electronics Engineering*

IMPATT DIODES AND CIRCUITS

IMPATT (impact avalanche transit time) diodes have the highest output power of any solid state device at millimeter-wave frequencies. Significant advances have been made in recent years to power combine these devices efficiently, and IMPATT diode transmitters can now rival the output power per unit volume of vacuum tube transmitters. IMPATT transmitters have several advantages over tubes, particularly for missile radar applications, so very high-power transmitters are being developed for several military systems. IMPATT diodes are also suited for low-cost commercial applications that only require the power of a few devices.

This article will explain the basic principles for designing IMPATT diode circuits, beginning with the characteristics of the diodes. General circuit principles will be discussed, and calculations of amplifier circuit performance will be illustrated. The general principles then will be applied to several types of single-diode circuits. Last, an approach for building very high-power transmitters with hundreds of diodes will be shown.

Much of the material in this article is drawn from the author's personal experience in designing IMPATT transmitters at Raytheon Systems Company. Many of the diodes and circuits that will be used for illustrations have military applications, so exact performance levels and design details cannot be revealed. However, this article will provide all the fundamental concepts and guidelines to design IMPATT transmitters for many applications.

Impatt Diode Overview

As a two-terminal device, an IMPATT diode generates radio frequency (*RF*) power by causing the RF current to be approximately 180° out of phase with the RF voltage (1). This is accomplished by two current delaying mechanisms, avalanche multiplication and transit time delay, which are the basis for the acronym IMPATT.

Figure 1 shows the doping profile $N(x)$ and electric field profile $E(x)$ of the most common type of IMPATT diode. The two curves are related by Poisson's equation in one dimension

$$dE/dx = -qN(x)/\epsilon$$

The integral of the electric field over distance (i.e, the area under the curve) is equal to the applied voltage plus the built-in diode potential.

The thin region in the center of the device is called the avalanche zone. The field is highest there, and avalanche multiplication occurs when a sufficient bias voltage is applied. To the right of the avalanche zone is a thin spike of n-type doping, which causes a large drop in the electric field. On the left side of the avalanche zone is a p-type doping spike that causes a similar drop in the electric field. These doping spikes confine the avalanche process to the thin, high-field avalanche zone.

The IMPATT structure in Fig. 1 also has an n-type drift zone on the right side of the avalanche zone and a p-type drift zone to the left. The doping in these regions is chosen to cause the electric field to slope gradually away from the center of the device. The right side of the device is bounded by a heavily doped n-type contact layer, and the left side is bounded by a p-type contact.

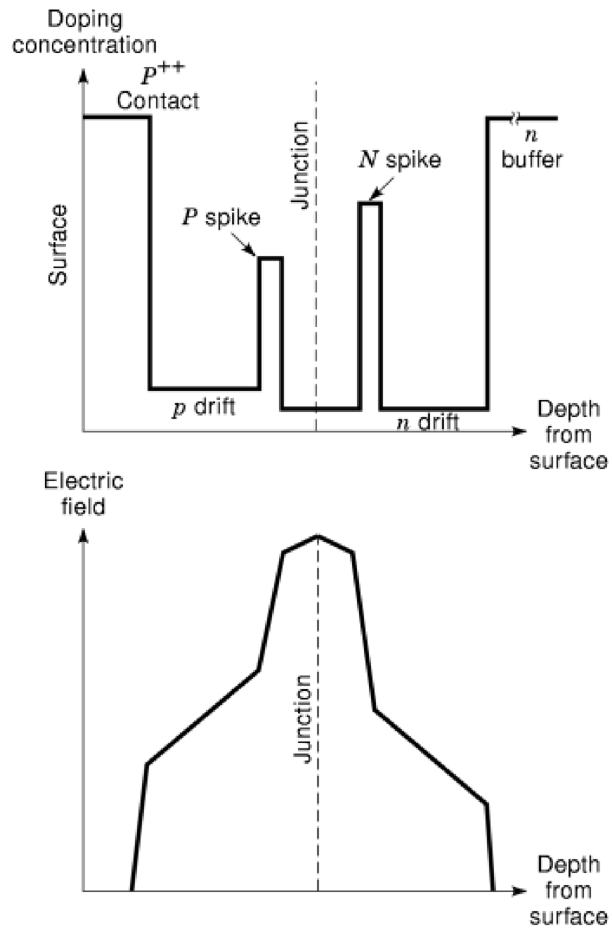

Fig. 1. Doping versus depth and electric field versus depth of a double-drift IMPATT diode.

Figure 2 explains how this structure accomplishes the necessary phase delay between the RF voltage and the RF current. The upper graph shows a sinusoidal RF voltage (35 GHz is used in this example) which is superimposed on a dc bias. The second graph shows the current that leaves the avalanche zone in response to the sinusoidal terminal voltage. Notice that the peak of the avalanche current is delayed by 90° from the peak of the terminal voltage. This phase shift occurs because the avalanching process needs many collisions, each with a very short delay, to generate a significant current. There are few electrons and holes at the beginning of the cycle, and most of the positive half of the RF cycle is needed for the current to grow exponentially to a significant level. In the second half of the cycle, the electric field is insufficient to sustain the avalanche, and the current quickly decays until the next cycle.

The pulses of electrons and holes that were generated in the avalanche zone during the positive half of the RF cycle flow with constant velocity through the drift zones during the negative half of the cycle. The motion of the large bunch of electrons toward the n-contact on the right and the motion of the holes to the left induce the external current shown in the third graph of Fig. 2. Notice that the terminal current is flowing primarily during the negative half-cycle of the terminal voltage (i.e., the diode behaves as a negative resistance at its operating frequency). This causes the IMPATT diode to deliver energy to the external circuit at the frequency

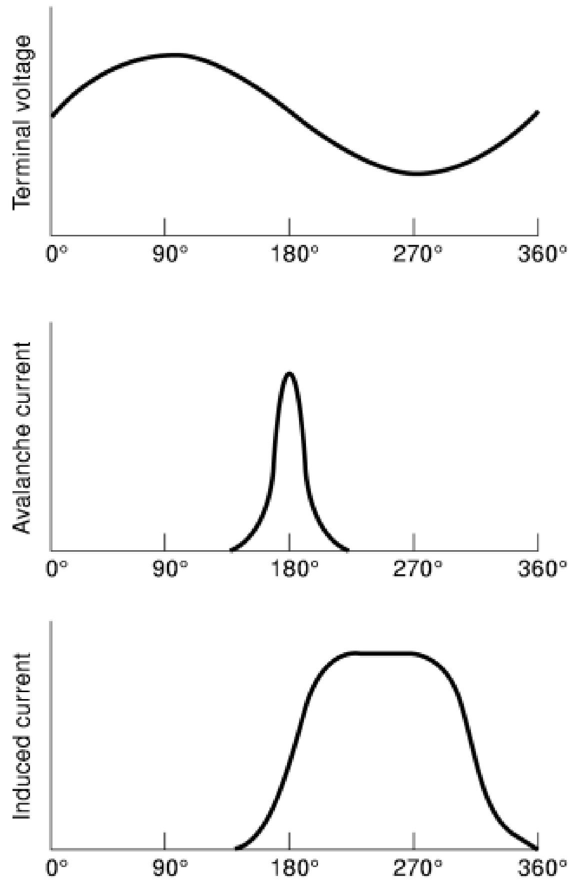

Fig. 2. Simplified theory of an IMPATT diode.

of the voltage excitation. The remainder of the bias power that is not converted to RF energy is dissipated in the diode as heat.

The concept of negative resistance is very useful to the circuit designer who must tailor the impedance of the IMPATT circuit to the impedance of the IMPATT diode to achieve the desired oscillation or amplification. However, to those readers who are not accustomed to this concept, another way to understand the power generation of an IMPATT diode is to consider the power dissipation in the diode relative to the applied dc power. The dc input power is the average dc voltage multiplied by the average dc current, whereas the power dissipation is the time average of the product of the instantaneous voltage and instantaneous current. These quantities are illustrated in the top and bottom graphs in Fig. 2. The dissipation is about 80% of the dc input power in this illustration, and the remaining 20% emerges from the device as RF power at the frequency of the RF voltage excitation.

Notice in Fig. 2 that the induced current waveform is not centered at 270° where the minimum RF voltage occurs. The length of the induced current pulse can be adjusted by changing the lengths of the drift zones, and the highest efficiency occurs when the induced current pulse is somewhat shorter than 180°.

It is important for a circuit designer to understand that the impedance of an IMPATT diode is a function of the RF voltage amplitude. Figure 3 illustrates the terminal voltage and induced terminal current at three different voltage amplitudes. The upper graph shows the small signal waveform conditions where both the

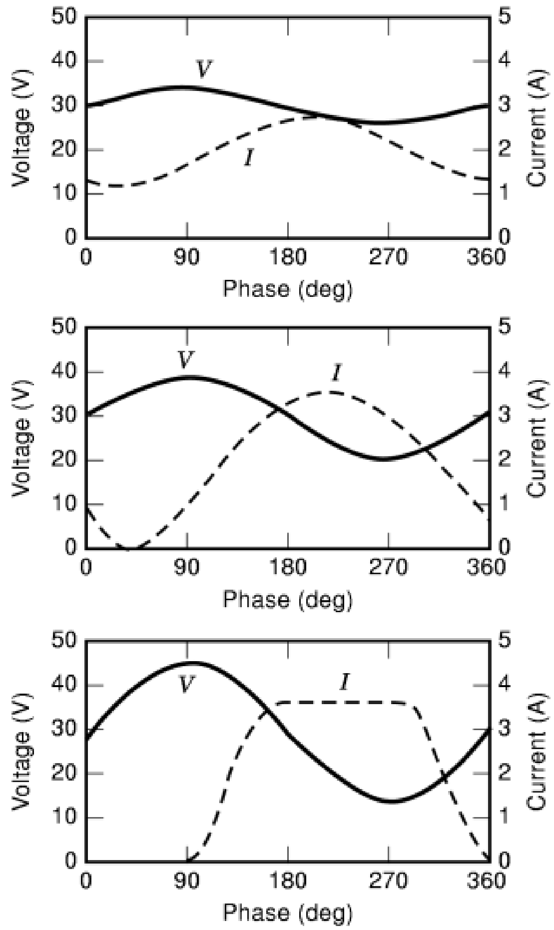

Fig. 3. Induced current in an IMPATT diode for low-, medium-, and high-voltage amplitudes.

current and the voltage are sinusoidal. The amplitude of the RF current is proportional to the RF voltage, causing the device impedance (the ratio of the RF voltage to the RF current) to be insensitive to the voltage amplitude. With the medium voltage amplitude shown in the center graph, the current waveform has become distorted, and additional increases in RF voltage will not result in proportional increases in the RF current. Then, under the large signal conditions shown in the lower graph, the current waveform is nearly a square wave. Additional increases in the RF voltage amplitude cannot cause additional increases in the amplitude of the induced current because the time average of the current waveform must equal the fixed bias current.

The admittance of the diode at its terminals can be extracted from these waveforms by computing the amplitude of the first harmonic of the current waveform and its phase relative to the RF voltage. Figure 4 shows the conductance and susceptance at the terminals of a 35 GHz IMPATT diode over the range of RF voltages shown in Fig. 3. The susceptance includes the fixed capacitive susceptance of the diode as well as the portion caused by the currents in Fig. 3.

Several observations about these curves are important to an RF circuit designer. Notice that negative conductance of the diode is considerably lower for high RF voltage amplitudes than for small signals. This will cause gain compression in IMPATT amplifiers. Notice also that the susceptance of the diode is a function of RF

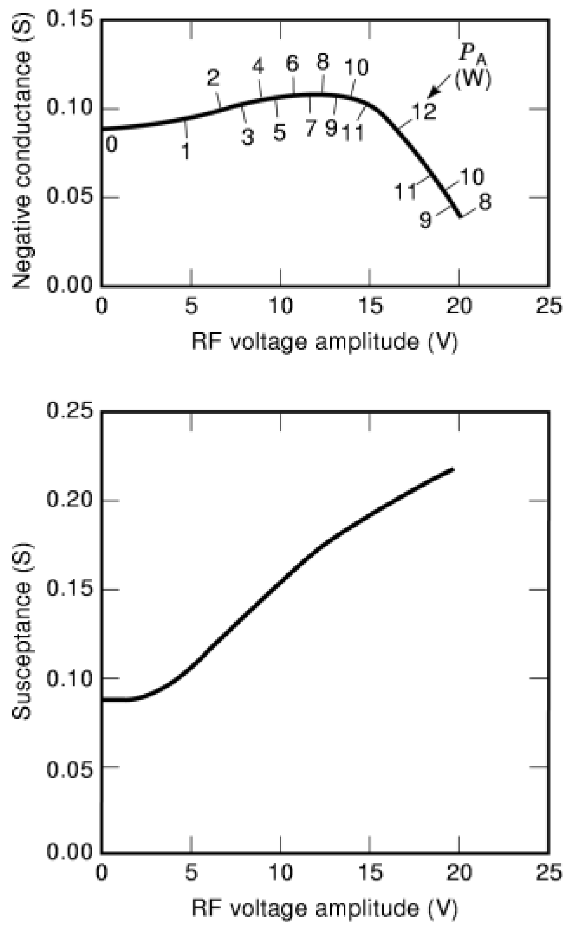

Fig. 4. IMPATT diode conductance and susceptance versus RF voltage amplitude.

voltage amplitude. This will cause the center frequency of an IMPATT amplifier to decrease as the RF signal level increases. It also causes the gain versus frequency characteristic of an IMPATT amplifier to drop more rapidly below the center frequency than on the high side of the center frequency. All these observations will be discussed later in this article.

The power P that is generated by an IMPATT diode can be calculated using

$$P = -(1/2)GV^2$$

where G is the chip conductance (a negative quantity) and V is the RF voltage amplitude.

Figure 4 has tick marks along the conductance versus RF voltage amplitude curve in intervals of 1 W. In this example, the maximum RF power is 12 W at an RF voltage of 16.5 V. At higher RF voltage amplitudes, other physical effects begin to affect the induced current waveform, and the output decreases.

A convenient way for the circuit designer to plot this information is to convert the chip conductance and susceptance data at each tick mark to chip impedance $-Z_C$ as illustrated in Fig. 5. Notice that the negative resistance of an IMPATT diode is relatively low, varying in this case from a small signal value of $-6.4\ \Omega$ to

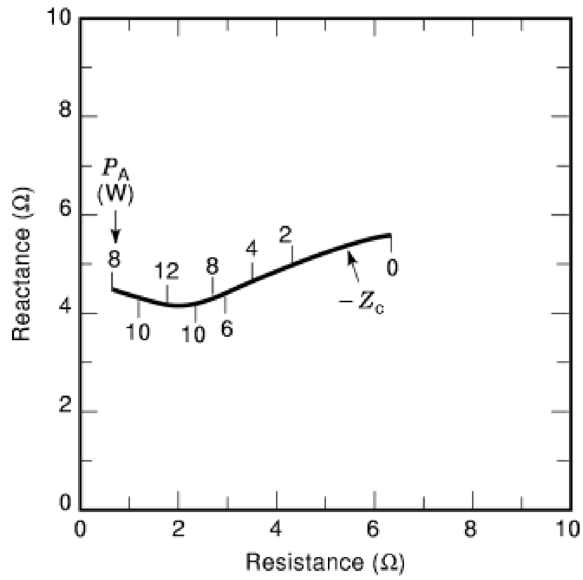

Fig. 5. IMPATT diode resistance and reactance as functions of RF power.

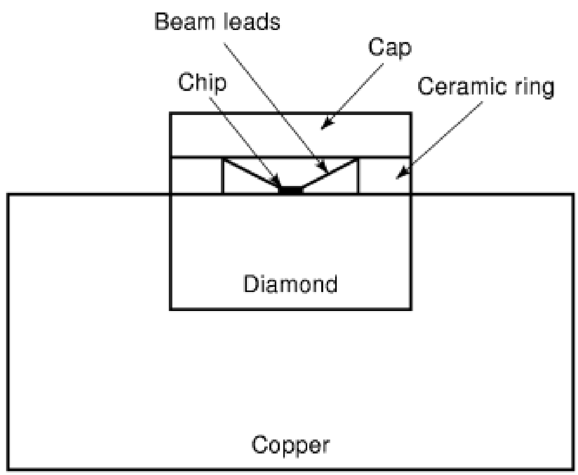

Fig. 6. Cross section of an IMPATT diode package.

a large signal value of −0.8 Ω. It is difficult to match low impedance levels like this over a broad frequency range, so IMPATT amplifiers are generally used for systems with fractional bandwidths of 5% or less. Most transmitters require less than 5% bandwidth, so this is not a significant restriction.

The electrical characteristics shown in Figs. 4 and 5 are approximations to the properties of a pulsed Ka-band IMPATT diode. Exact values have not been used because these devices are useful in military systems. The shapes of the curves are qualitatively correct, and they will be used in later sections to explain circuit design principles.

Figure 6 is a cross-sectional drawing of a typical millimeter-wave IMPATT diode package. The chip is thermal compression bonded to a diamond heat-sink, which gives the best possible heat-sinking. The diamond

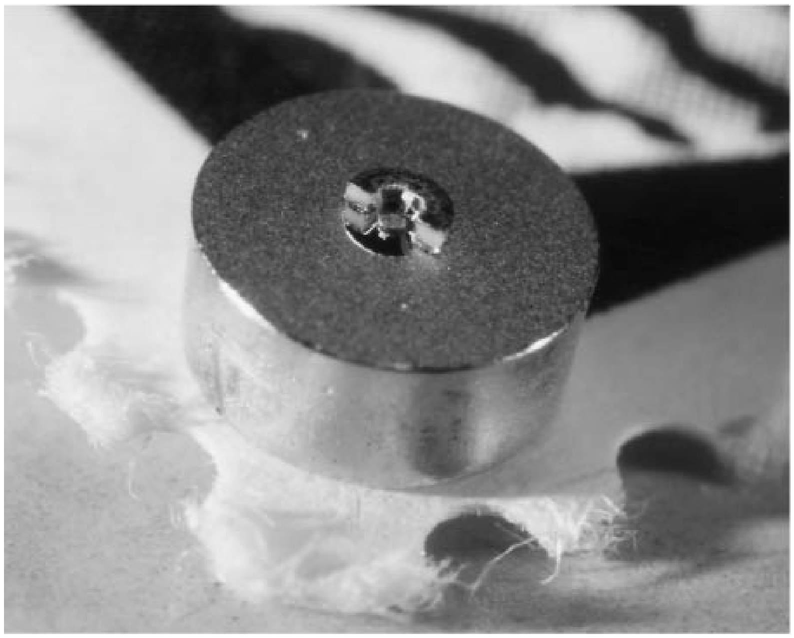

Fig. 7. Photograph of an IMPATT diode package.

is imbedded in a copper puck to facilitate a good thermal contact to the bottom of the diode package. The chip is surrounded by a ceramic or quartz insulating ring, and beam leads or bonding straps connect the top of the chip to the top of the insulator. A cap is soldered over the top of the ring to provide mechanical protection and an hermetic enclosure.

Figure 7 is a photograph of a Raytheon device that has one-half of the ceramic ring removed to reveal the internal construction. This type of package can be easily inserted or removed from an RF circuit, being held in place with a copper screw for heat removal. The ring is as small as possible to minimize the electrical transformation through the parasitic inductances and capacitances of the package. Low-cost IMPATT diodes at lower microwave frequencies have a threaded stud in place of the diamond puck.

Impedance Matching of a Negative Resistance Device

An IMPATT diode can be used as a stable amplifier, an injection-locked oscillator or a free-running oscillator. The mode of operation is determined by the load impedance that the RF circuit presents to the IMPATT chip. This section will explain the general principles of IMPATT diode impedance matching to lay a foundation for later discussions of specific circuits.

Free-Running Oscillator. Figure 8 contains the IMPATT chip impedance curve $-Z_C$ from Fig. 5 (the "device line") plus a curve of load impedance versus frequency Z_L (the "load line"). In Fig. 8, we will assume that the impedance of this 35 GHz diode does not change appreciably over a frequency range of 2 GHz (a 6% bandwidth). The load line of the circuit Z_L crosses the device line $-Z_C$ at 35 GHz near the highest power point on the device line. At the intersection, the condition for oscillation is met

$$Z_C + Z_L = 0$$

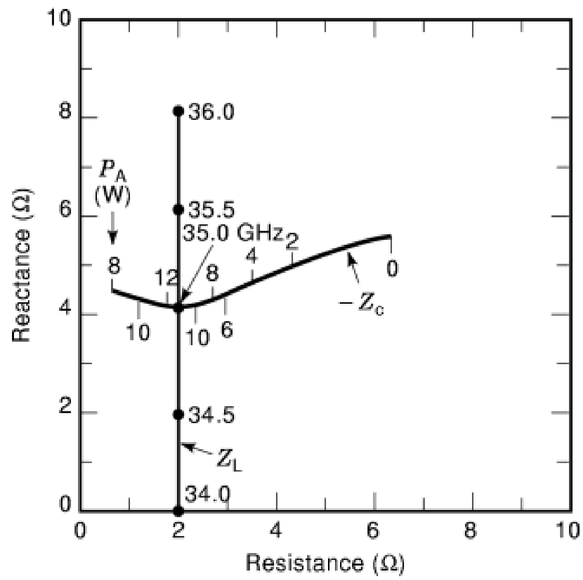

Fig. 8. IMPATT device line and load line for an oscillator.

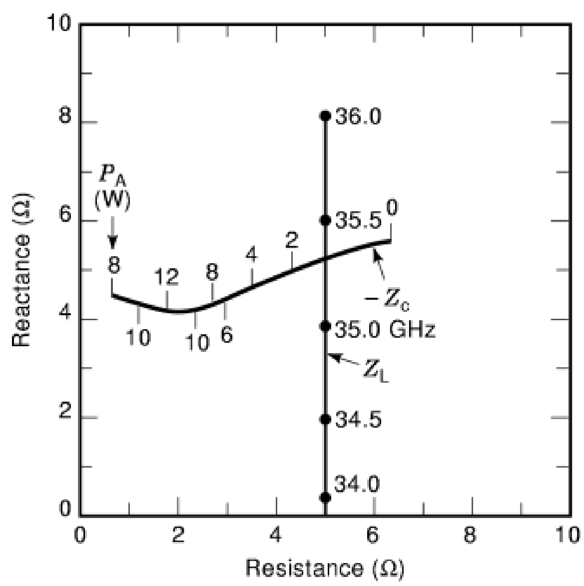

Fig. 9. IMPATT device line and load line for an injection-locked oscillator.

If the load line is moved vertically, the frequency of intersection will shift, causing the diode to oscillate at a new frequency. If the load line is moved horizontally, the intersection will occur at a different RF output power level. Moving the load line horizontally also changes the frequency slightly as a result of the slope of the device line.

Injection-Locked Oscillator. Figure 9 is similar to Fig. 8, but the load line is shifted to the right. The point of intersection occurs at 35.25 GHz on the load line and at less than full power on the device line. In the absence of an RF input signal, the diode will oscillate in this circuit at 35.25 GHz with an output power of about 1 W.

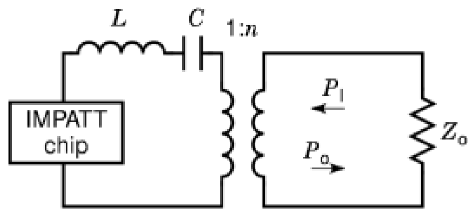

Fig. 10. Equivalent circuit for a one-diode IMPATT amplifier.

If a 35.25 GHz RF input signal is applied, the power P_A that is added by the diode will combine with the RF input signal P_I to produce an RF output signal P_O that is larger than the input signal

$$P_O = P_I + P_A$$

The gain G of the circuit is the ratio of RF output to RF input

$$G = P_O / P_I$$

These two equations can be combined to solve for both P_O or P_I in terms of P_A and G

$$P_I = P_A/(G - 1) \qquad (1)$$

$$P_O = GP_A/(G - 1) \qquad (2)$$

Figure 10 illustrates a simple equivalent circuit for an IMPATT amplifier. A more accurate approach to circuit characterization will be discussed later, but the basic principles can be explained more clearly with this simple model. The model is an LC circuit in series with the IMPATT chip Z_C. The coupling between the resonant circuit and the external load Z_0 is an ideal transformer with a turns ratio of n.

The input impedance Z at angular frequency ω of the diode and circuit shown in Fig. 10 is

$$Z = n^2[Z_C + j\omega L + 1/(j\omega C)]$$

This equation can be further simplified by substituting $j\omega X_L = j\omega L + 1/(j\omega C)$, so that

$$Z = n^2(Z_C + j\omega X_L) \qquad (3)$$

The voltage reflection coefficient Γ from an impedance Z in a transmission line with characteristic impedance Z_0 is

$$\Gamma = (Z - Z_0)/(Z + Z_0) \qquad (4)$$

Substituting Eq. (3) into Eq. (4) and making a few algebraic steps,

$$\Gamma = [n^2(Z_C + j\omega X_L) - Z_0]/[n^2(Z_C + j\omega X_L) + Z_0]$$
$$\Gamma = [(Z_C + j\omega X_L) - Z_0/n^2]/[(Z_C + j\omega X_L) + Z_0/n^2]$$
$$\Gamma = [Z_C - (Z_0/n^2 - j\omega X_L)]/[X_C + (Z_0/n^2 + j\omega X_L)] \qquad (5)$$

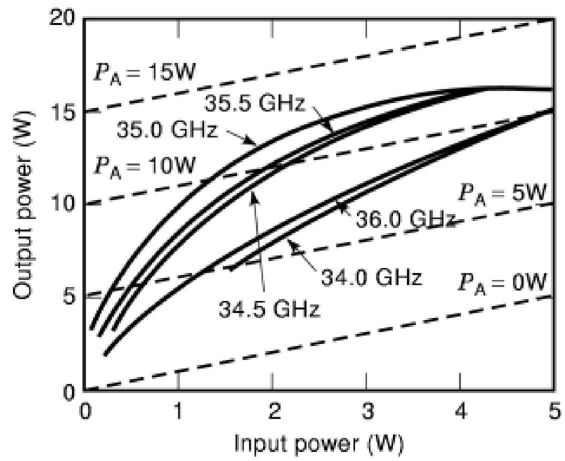

Fig. 11. RF output power versus RF input power of an injection-locked oscillator.

The load impedance seen by the diode is $Z_L = (Z_0/n^2) + j\omega X_L$ and its complex conjugate is $Z^*{}_L = (Z_0/n^2) - j\omega X_L$. Substituting these quantities into Eq. (5),

$$\Gamma = (Z_C - Z_L^*)/(Z_C + Z_L) \tag{6}$$

The gain of the amplifier is the square of the magnitude of the reflection coefficient

$$G = |\Gamma|^2 = |(Z_C - Z_L^*)/(Z_C + Z_L)|^2 \tag{7}$$

Equations (1), (2), and (7) can be used to construct curves of P_O versus P_I from the data in Fig. 9 using the following procedure.

(1) Choose a value of $-Z_C$ on the device line (start at the left end of the device line) and note the added power P_A.
(2) Note the value of Z_L from the load line at the frequency of operation.
(3) Calculate the gain G using Eq. (7).
(4) Calculate the RF input power P_I and RF output power P_O using Eqs. (1) and (2).
(5) Repeat this calculation for several other values of $-Z_C$ and draw a curve of P_O versus P_I.

As the choice for $-Z_C$ is moved to the right along the device line toward the intersection of the device line and the load line, the gain of the circuit increases, and the calculated input power decreases. When $-Z_C$ is chosen to be the intersection between the device line and the load line, the gain is infinite, and the RF input power is zero, which is the free-running oscillator condition. Values of $-Z_C$ that are to the right of the intersection point will not result in stable operation, and the injection-locked oscillator will "break lock" (2) from the RF input signal and generate a free-running signal.

Figure 11 shows curves of P_O versus P_I at five different frequencies. All the curves were calculated with the same procedure, but with different values of Z_L from the load line for each of the frequencies. Notice that the curves do not pass through the origin because of the free-running oscillation at 35.25 GHz. At frequencies near 35.25 GHz, P_O increases rapidly for small values of P_I because the gain is very high when $-Z_C$ is close

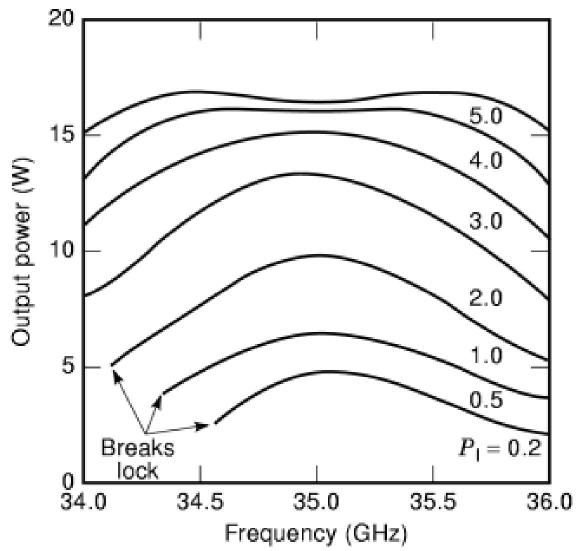

Fig. 12. RF output power versus frequency of an injection-locked oscillator.

to Z_L [see Eq. (7)]. For larger values of P_I, P_O continues to increase but at a slower rate because the gain is steadily decreasing. This effect is called gain compression.

The diagonal lines in Fig. 11 are contours of constant added power. At 35.0 GHz, the maximum added power is 12 W, which occurs at an input power of 3 W. For maximum efficiency, the circuit should be designed to operate near this point. Notice, however, that the added power of the circuit is relatively constant over a modest range of input powers. This makes IMPATT circuits tolerant to being underdriven or overdriven.

The data in Fig. 11 shows how an injection-locked IMPATT circuit behaves as a function of RF input power at several fixed frequencies, and these are called drive curves. Data points from this graph can be replotted such that the RF input power is held constant while the frequency is varied. This is illustrated in Fig. 12, and these are the frequency response curves of the IMPATT circuit. There are several interesting observations regarding these curves. First, the curves are not symmetrical because the device line is tilted relative to the load line in Fig. 9. As the RF input drive is decreased, the circuit becomes unlocked at the lower edge of the operating frequency range before it becomes unlocked at the high end of the band.

Another important observation is that the bandwidth of the circuit is greater when the RF input power is higher. When the RF input power is 0.2 W, the gain is high at 35.0 GHz, but it decreases rapidly for other frequencies. When the RF input power is between 3 and 4 W at 35.0 GHz, the circuit adds about 12 W. At 5 W of RF input, the diode is overdriven and only 11 W are added. This saturation behavior causes the output power and gain to be relatively constant over the entire frequency range at the higher drive levels.

Stable Amplifier. The mode of operation in Figs. 9, 11, and 12 is called injection locked, and the circuit is called an injection-locked oscillator. The gain of an injection-locked oscillator as illustrated in Figs. 11 and 12 will increase if the circuit designer moves the device line in Fig. 9 to the left toward the maximum added power point along the device line. This will be accompanied by a corresponding decrease in injection-locking bandwidth. Conversely, the gain will decrease and the bandwidth will increase, if the load line is moved to the right in Fig. 9. If the load line is moved far enough to the right, there is no intersection between the load line and the device line. In this case, the IMPATT circuit will not oscillate if the RF input signal is removed. This type of circuit is called a stable amplifier.

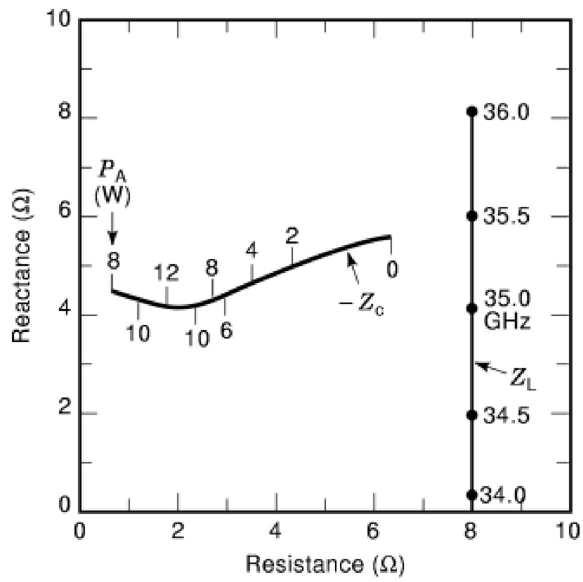

Fig. 13. IMPATT device line and load line for a stable amplifier.

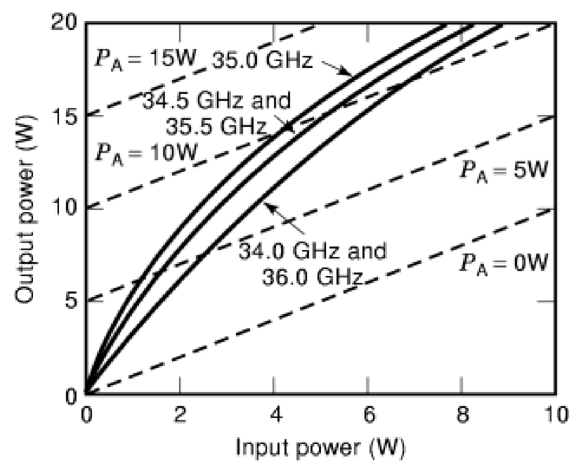

Fig. 14. RF output power versus RF input power of a stable amplifier.

Figure 13 illustrates relationship between the device line and the load line for a stable amplifier. The same procedure that was used earlier to calculate the drive curves and frequency response curves can be applied to the stable amplifier case. Figures 14 and 15 illustrate the resulting graphs. Notice that all the curves in Fig. 14 begin at the origin because the output of a stable amplifier is always "locked" to the RF input signal. Both figures show that the gain of the circuit is lower, and Fig. 15 clearly demonstrates an accompanying increase in bandwidth.

The choice of operating mode—oscillator, injection-locked oscillator, or stable amplifier—depends on the application. A free-running oscillator is useful for generating a high-power RF signal with a minimum of hardware. For example, a mechanically tuned oscillator is a good source of RF power for laboratory instrumentation.

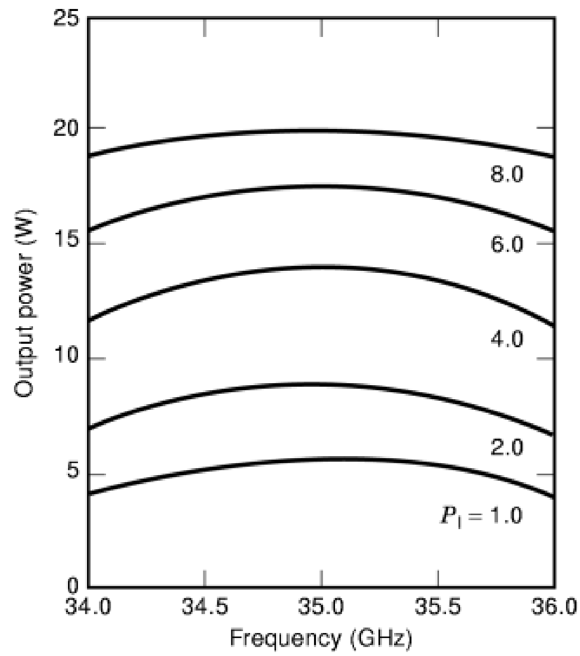

Fig. 15. RF output power versus frequency of a stable amplifier.

An injection-locked oscillator is the best for most transmitter applications. Gains of 3 dB to 10 dB are typical, and multiple stages can be cascaded for higher overall gain. If the bandwidth requirement is small, fewer stages with higher gain per stage can be used. The drive curve of an injection-locked oscillator is very nonlinear (see Fig. 11), and a multiple-stage transmitter will have many decibels of gain compression. Therefore, injection-locked oscillators are used only with frequency-modulated or phase-modulated signals.

An important application for injection-locked IMPATT amplifiers is pulsed radar systems. The bias current to the IMPATT diodes is turned on and off with the same waveform as the RF input signal. In this way, the IMPATT amplifier does not generate a free-running signal or noise while the radar receiver is detecting the reflected signal from the target. Pulsing the diode bias also reduces the prime power requirement by the duty cycle of the RF waveform.

Stable amplifiers are used when a free-running oscillation cannot be tolerated. A stable IMPATT amplifier might also be chosen for an amplitude-modulated signal if a modest amount of gain compression is acceptable. The use of constant voltage bias instead of constant current bias can be used to produce more linear drive curves (3), but constant voltage bias can also produce parametric instabilities, which will be discussed in a later section.

Effect of Load Mismatch. One final note of caution for the circuit designer is to provide a good external load for the IMPATT amplifier circuit. In the preceding examples, a perfect matched load Z_0 was assumed. A load impedance other than Z_0 will perturb the circuit impedance seen by the diode, changing the response of the amplifier. In practice, an amplifier circuit will be connected to a circulator to separate the output and input signals, and circulators have a return loss of about −25 dB over a 5% bandwidth.

A convenient way to estimate the effect of load mismatch is to estimate the amplitude of the wave that is reflected from the mismatch. This signal will combine with the intended RF input signal, resulting in a different amplitude. The maximum combined amplitude will occur when the voltage vectors for the two signals are in phase, and the minimum occurs when they are out of phase. The engineer should estimate the maximum and minimum voltage vectors and then calculate the maximum and minimum RF input power to the amplifier circuit.

For example, if an amplifier circuit has a gain of 5 dB and the load to which the circuit is connected has a return loss of 25 dB, the portion of the amplifier output signal which is reflected from the load will be 20 dB (a factor of 0.01) below the intended or nominal RF input signal. The voltage amplitude of a signal is proportional to the square root of its power, so the voltage amplitude of the signal that is reflected from the load will be approximately 0.10 relative to the voltage amplitude of the nominal RF input signal. The maximum RF input occurs when the nominal input signal and the signal that is reflected from the load are in phase

$$\text{Max}\, P_\text{I} \approx (1+0.10)^2\, \text{Nominal}\, P_\text{I} = 1.21\, \text{Nominal}\, P_\text{I}$$

The minimum RF input occurs when the two signals are out of phase

$$\text{Min}\, P_\text{I} \approx (1-0.10)^2\, \text{Nominal}\, P_\text{I} = 0.81\, \text{Nominal}\, P_\text{I}$$

It can be seen that even a good external load causes a modest uncertainty in the RF drive to an amplifier circuit. Fortunately, the gain compression that is evident in the drive curves in Figs. 11 and 14 causes the added power of a high-power IMPATT amplifier to be insensitive to an input power range of this size. The early stages of a multiple-stage amplifier will be more sensitive to load mismatch, however, because the gains are usually higher, and there is less gain compression when the diodes are operated at lower power levels.

Parametric Oscillation

The preceding sections of this article have explained how an IMPATT diode generates negative resistance at the frequency of operation and how the impedance of the circuit must be tailored to the device impedance. An IMPATT diode is very nonlinear when the RF voltage is high, and parametric effects cause it to develop negative resistance over a wide frequency range. This section will explain the origin of the parametric negative resistance and how to design an IMPATT circuit to prevent unwanted oscillations.

Bias Circuit Oscillation. One source of instability in IMPATT diodes is caused by the decreased power dissipation that occurs when the RF voltage across the diode is increased and the device generates more RF power. The ionization rates in the avalanche zone are higher if the temperature is reduced, so the dc bias current will increase if the voltage is held constant. A run-away condition can occur in an IMPATT oscillator because an increase in the bias current will cause a further increase in the RF voltage amplitude. This type of instability can cause oscillation in the bias input to an IMPATT amplifier or oscillator at frequencies up to about 10 MHz. The effect diminishes at higher frequencies because the thermal time constant of the IMPATT chip and package are too long to allow the junction temperature to respond to the changes in dissipated power.

Preventing this form of bias oscillation requires a high-impedance bias current source. When the RF voltage increases incrementally and the dissipated power decreases, the bias voltage will drop incrementally if the current is held constant. The RF properties of the diode are relatively insensitive to small changes in the operating voltage, so there is not a positive feedback effect as there is in the case of constant voltage bias. The drop in operating voltage that is caused by an increase in RF voltage amplitude is called the rectification effect or back bias effect.

The simplest way to create a high impedance dc bias source in the laboratory is to insert a noninductive power resistor between the bias input of the IMPATT circuit and a dc power supply. The minimum value of resistance that is needed for stability will need to be determined experimentally, but 100 Ω is a good starting value. The resistor should consist of a network of lower power, noninductive resistors rather than a single, high-power, wire-wound resistor. The 'constant current' mode of a power supply should not be used because the response time of the feedback loop within the supply is much too long to protect the diode from a runaway condition.

The primary disadvantage to using a series resistor to make a high-impedance current source is the amount of heat that is dissipated in the resistor. A better approach is to design a constant current electronic bias circuit with a low-voltage drop (typically 5 V). If a *pnp* transistor or a *p*-channel FET is used for the pass transistor in the bias circuit, the bias output will come from the collector or drain and have a relatively high impedance well beyond 10 MHz. If the IMPATT circuit is to be used in a pulsed radar transmitter, the electronic bias circuit must be designed for both high output impedance and fast switching speed.

Another effect occurs at high RF voltage amplitudes when the current leaving the avalanche zone becomes very low between cycles. In IMPATT diodes that are designed for microwave frequencies such as 10 GHz, the avalanche current waveform may reach the saturation current level. If the RF voltage amplitude is increased further, the avalanche current cannot drop below the saturation current level. During the positive half of the RF voltage cycle, the avalanche current builds up too quickly from this lower bound. This will result in a higher peak avalanche current and a higher dc current if the bias voltage is not reduced. This effect is quite strong in low-frequency devices, and it generates negative resistance to several gigahertz. However, this mechanism does not exist in millimeter-wave diodes because there is insufficient time between RF cycles for the avalanche current to drop to the ionization current level.

An IMPATT bias circuit must have a high impedance to several gigahertz to prevent the last type of parametric oscillation. It is particularly important to have low shunt capacitance in the bias circuit of the IMPATT circuit and in the bias supply. There are several comprehensive articles on this subject from the early years of IMPATT research (4,5). The principles in these articles can be adapted to other types of IMPATT circuits.

Subharmonic Oscillation. Another type of parametric effect that is common in IMPATT circuits is subharmonic oscillation (6,7,8,9). Simply stated, the IMPATT diode generates broad band negative resistance as a result of the nonlinear behavior of the avalanche process. One form of subharmonic oscillation is a signal at exactly one-half of the fundamental operating frequency. Subharmonic oscillation can also appear as a pair of signals whose frequencies add up to the fundamental frequency.

Subharmonic oscillation is very difficult to prevent because of the very broad frequency range where it can occur, so it is usually present in IMPATT amplifiers. A general approach to minimize the magnitude of subharmonic oscillation is to load the bias port of the IMPATT circuit with a microwave-absorbing material. This load will absorb some of the RF power at the fundamental, but the circuit designer can find creative ways to minimize the lost power.

If the circuit designer uses a high-impedance bias supply, minimizes the shunt capacitance in the bias circuit of the IMPATT circuit and in the bias supply, and also loads the bias port with a microwave absorber, the IMPATT circuit should behave well.

Types of Impatt Circuits

All the general principles of IMPATT diode operation and the RF circuit requirements have been discussed in the previous section of this article. This section will briefly illustrate several types of IMPATT circuits that meet these general requirements.

An IMPATT circuit must present a load resistance of only a few ohms to an IMPATT chip, so a large transformation is required between the characteristic impedance of the RF port of the IMPATT circuit and the chip. In a coaxial transmission line medium, this can be accomplished with a coaxial matching transformer as illustrated in Fig. 16. The low-impedance transformer is created by reducing the diameter of the outer conductor rather than increasing the diameter of the inner conductor because this results in lower parasitic inductance in the mount. The "puck" style of IMPATT package that was illustrated in Figs. 6 and 7 can be easily installed and removed from this kind of diode mount.

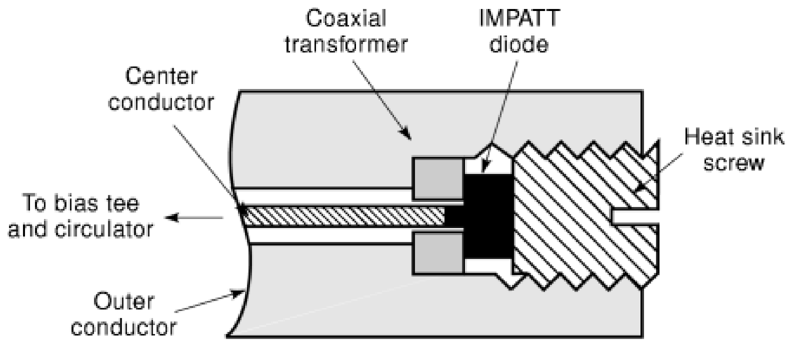

Fig. 16. Coaxial circuit.

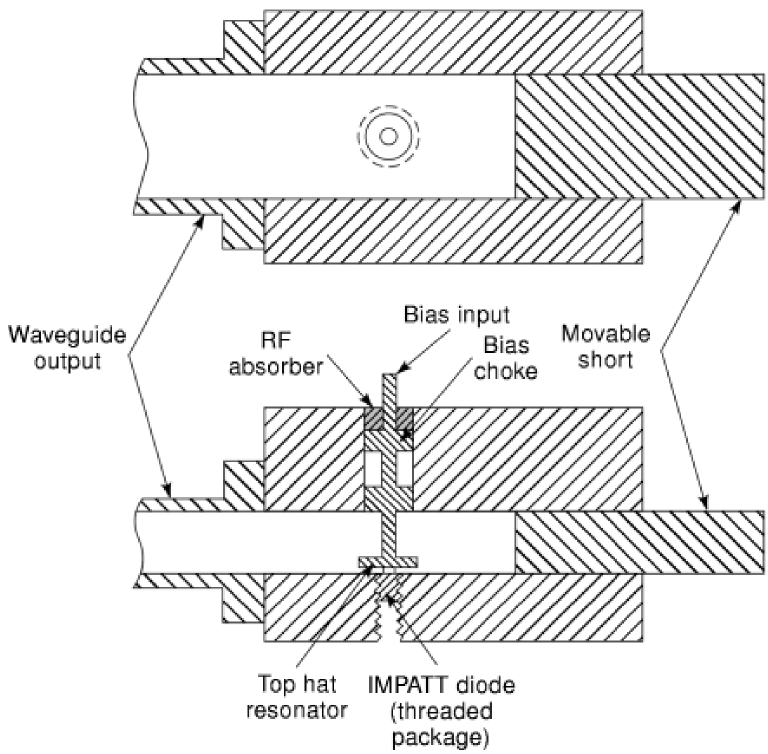

Fig. 17. Top hat circuit.

The matching transformer (typically 1/8 wavelength long because of the parasitic reactance of the package and the mount) is designed to transform the characteristic impedance of the coaxial line (typically 50 Ω) to the desired load impedance. Wider bandwidth can be achieved with a two-section transformer. Bias is applied to the center conductor through a bias tee (not shown), and the RF input and output signals are separated by a coaxial circulator (not shown). The circulator should have wide bandwidth to provide RF loading in the subharmonic frequency range.

The top hat circuit (15), illustrated in Fig. 17, was used extensively in the early years for oscillator testing of IMPATT diodes. The diode is mounted in the floor of a waveguide cavity (a threaded diode package works

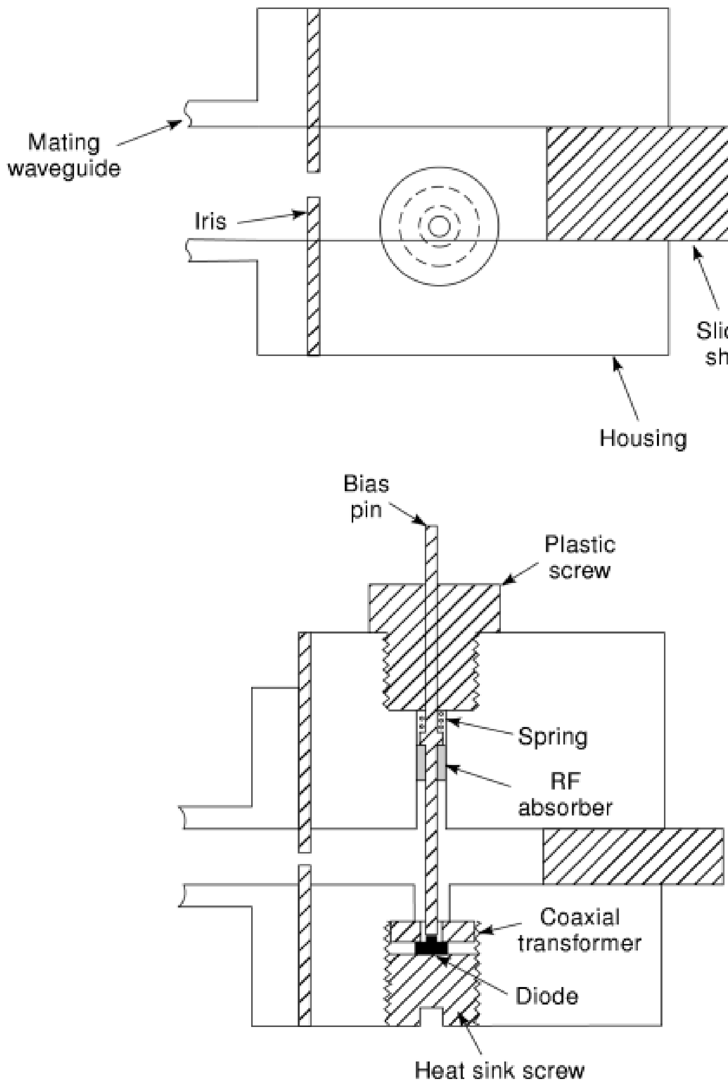

Fig. 18. One-diode rectangular Kurokawa circuit.

well with this type of circuit), and a top hat resonator is placed on top of the diode. The region under the top hat is a radial transmission line that transforms the high impedance of the waveguide to the low impedance of the chip. The frequency of oscillation is set primarily by the diameter of the top hat, and the backshort is adjusted to set the cavity resonant frequency to be equal to the top hat frequency. The load resistance that is seen by the IMPATT diode is set by the spacing between the top hat and the floor of the waveguide. At millimeter-wave frequencies, the diameter of the top hat might be as small as the diameter of the bias pin. An E-H tuner or slide-screw tuner can be connected to the waveguide output for fine-tuning. The bias pin contains an RF absorber for parametric stability.

A very common and useful type of IMPATT circuit is called the rectangular Kurokawa circuit (10) that is illustrated in Fig. 18. A resonant cavity is formed in a rectangular waveguide between the movable short

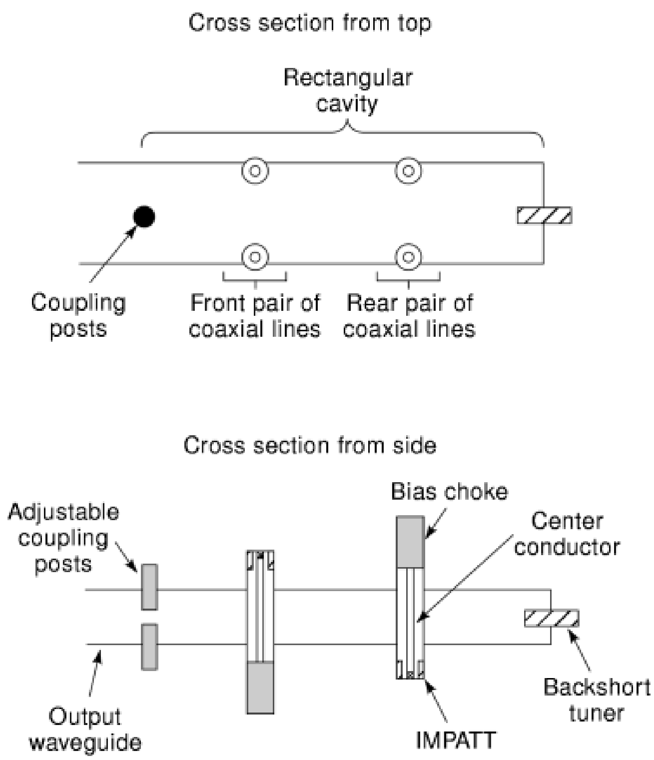

Fig. 19. Four-diode rectangular Kurokawa circuit.

circuit at the right and an iris on the left. A coaxial transmission line passes through the broad walls of the waveguide cavity to which it is magnetically coupled. An IMPATT diode and a matching transformer are located at one end of the coaxial line, and an RF absorber is inserted into the other end. The coaxial transformer is the primary impedance matching element, and the load resistance and reactance are controlled by the transformer impedance and length, respectively.

The cavity resonance in the rectangular Kurokawa circuit causes a significant perturbation in the diode load impedance near the operating frequency. The movable short is adjusted to make the cavity resonant frequency correspond to the resonant frequency of the diode and matching transformer, and the hole diameter in the iris controls the coupling to the external load. Far from the operating frequency, the cavity has little effect, and the RF absorber helps prevent parametric oscillation.

The Kurokawa circuit is very useful for diode testing because a wide range of load impedances can be achieved with different matching transformers and irises. The RF load in the coaxial line absorbs about 25% of the generated power, but the circuit efficiency can be measured with an automatic network analyzer so that the RF power of the diode can be calculated.

The Kurokawa circuit can be modified to be a power combiner (11) by adding additional coaxial lines as shown in Fig. 19. The pairs of coaxial lines are separated by one-half of a wavelength so that the coupling from the waveguide cavity to each coaxial line is approximately the same.

A variation of this last circuit type is the cylindrical Kurokawa circuit illustrated in Fig. 20. The coaxial lines that contain the diodes are the same as the ones in the rectangular Kurokawa, but the cavity is now cylindrical instead of rectangular. The RF output port is a coaxial line in the top of the cavity, and the coupling between the cavity and the external load is adjusted by changing the penetration of the electric field probe.

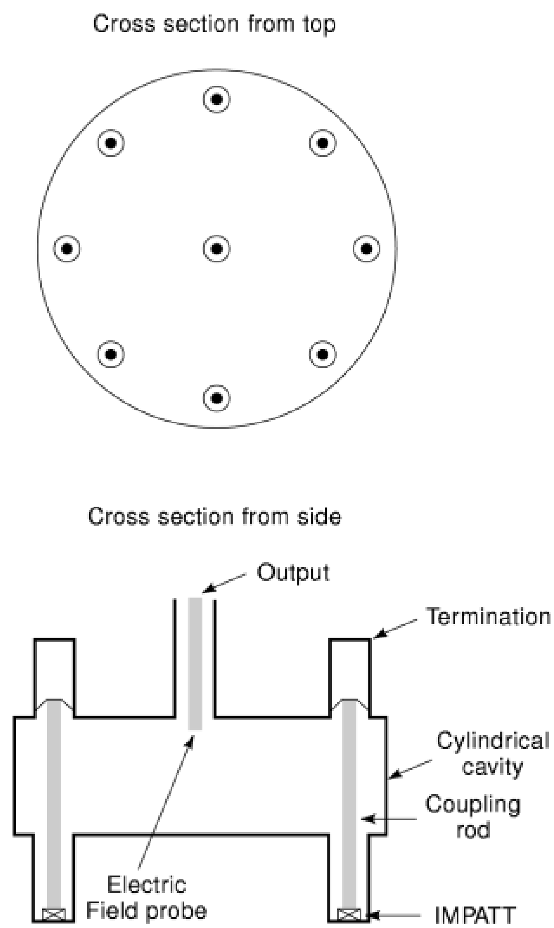

Fig. 20. Eight-diode cylindrical Kurokawa circuit.

Impatt Circuit Characterization

The discussion of impedance matching a negative resistance device used a lossless series resonant equivalent circuit. In practice, an IMPATT circuit is neither lossless nor does its impedance vary with frequency like this simple model. However, this model is qualitatively correct and useful for the purposes of illustration.

It is not practical to measure directly the load impedance that is seen by an IMPATT diode because the load impedance is only a few ohms and the chip location is inaccessible to a network analyzer. However, those IMPATT circuits where the diode is mounted in a coaxial transmission line can be characterized indirectly with a network analyzer. If the diode and its matching transformer are removed, a network analyzer with a coaxial input can be connected to the circuit. After the two-port S-parameters of the circuit are measured, an equivalent circuit for the matching transformer and the diode package parasitic reactance can be used to calculate the load impedance at the diode chip. Figure 21 illustrates how a network analyzer would be used to make this two-port measurement of a one-diode rectangular Kurokawa circuit.

There is another measurement technique that requires only one-port network analyzer measurements. Figure 22 shows a rectangular Kurokawa circuit with the diode and its coaxial transformer removed. These

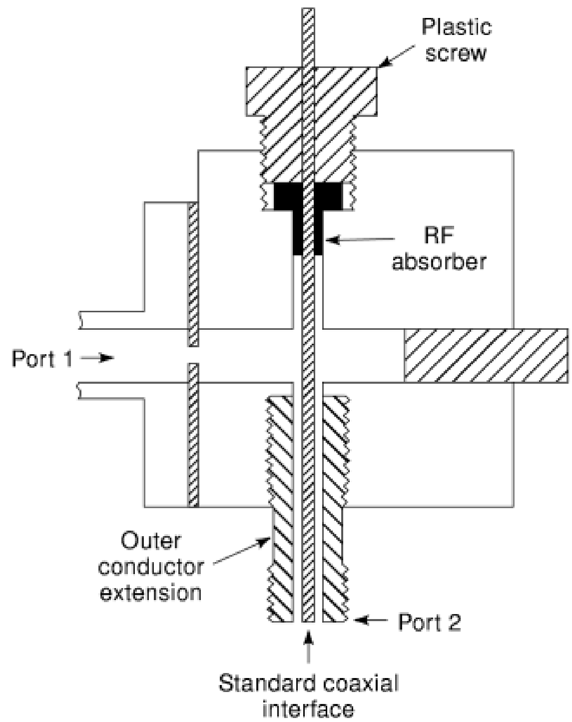

Fig. 21. Direct characterization of an IMPATT circuit with a coaxial adapter.

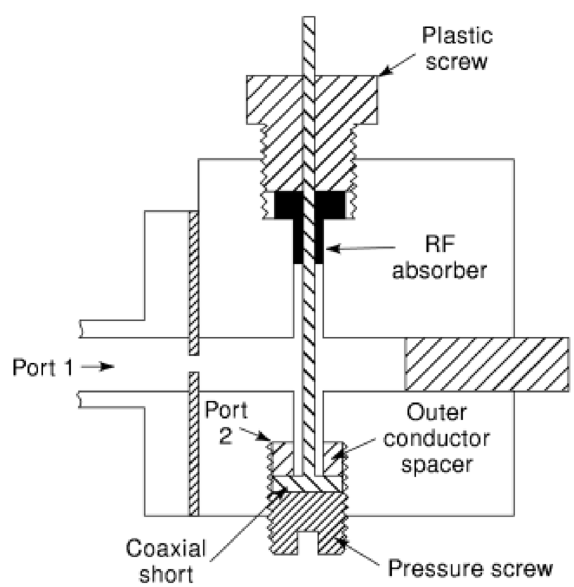

Fig. 22. Characterization of an IMPATT circuit by using offset short circuits.

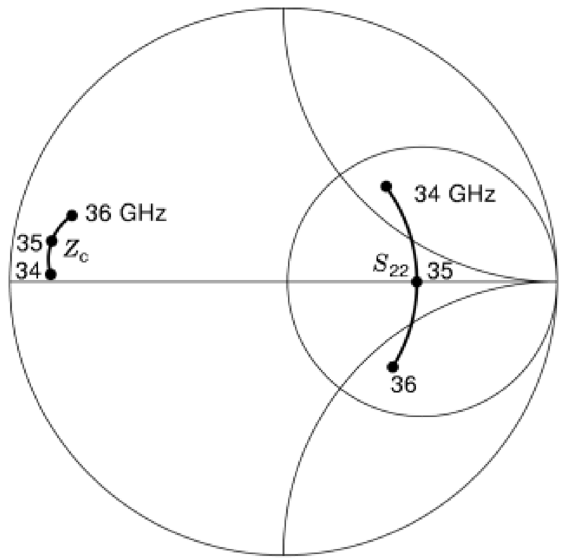

Fig. 23. Smith chart impedance plot of the S_{22} and chip load impedance of an oscillator circuit.

components are replaced by a coaxial short and an outer conductor spacer to create an offset short. The reflection coefficient Γ_2 of a load that is attached to port 2 is transformed to Γ_1 at port 1 by the S-parameters of the circuit according to the equation

$$\Gamma_1 = S_{11} + S_{21}S_{12}\Gamma_2/(1 - S_{22}\Gamma_2)$$

This equation has three unknowns (S_{11}, S_{22}, and the product $S_{21}S_{12}$) which fully describe the circuit at a single frequency. Different values of Γ_2 are created by three offset shorts, and Γ_1 is measured for each one. This gives three simultaneous equations that can be solved for the three unknowns. More accurate results over a wider range of circuit impedance can be obtained by using more than three (typically five to seven) offset shorts and using numerical methods to find a least-squares best fit for the three unknowns. The procedure to calculate the circuit S-parameters over a frequency range of interest is easily programmed into a computer that controls an automatic network analyzer.

Figure 23 illustrates the S_{22} of a one-diode rectangular Kurokawa oscillator circuit. The load impedance at the chip was calculated by referring the S_{22} data through an equivalent circuit for the matching transformer and package parasitic reactance. Notice that the load resistance at the chip is nearly constant and the reactance increases linearly with frequency. This impedance versus frequency behavior was the basis for choosing a series resonant equivalent circuit in the earlier discussion on impedance matching. The oscillator circuit efficiency η is the ratio of the power that emerges from the oscillator divided by the power generated by the diode

$$\eta = |S_{21}S_{12}|/(1 - |S_{22}|^2)$$

This relationship is used during IMPATT diode oscillator testing to calculate the RF output power of the diode.

The offset short method for characterizing a circuit has several advantages and disadvantages compared to a direct two-port network analyzer measurement. The offset method can be used to characterize a multiple diode combiner circuit if all the coaxial ports are electrically identical. To do this, the coaxial lines are loaded with identical offset shorts. Network analyzer measurements are taken for three or more sets of these offset

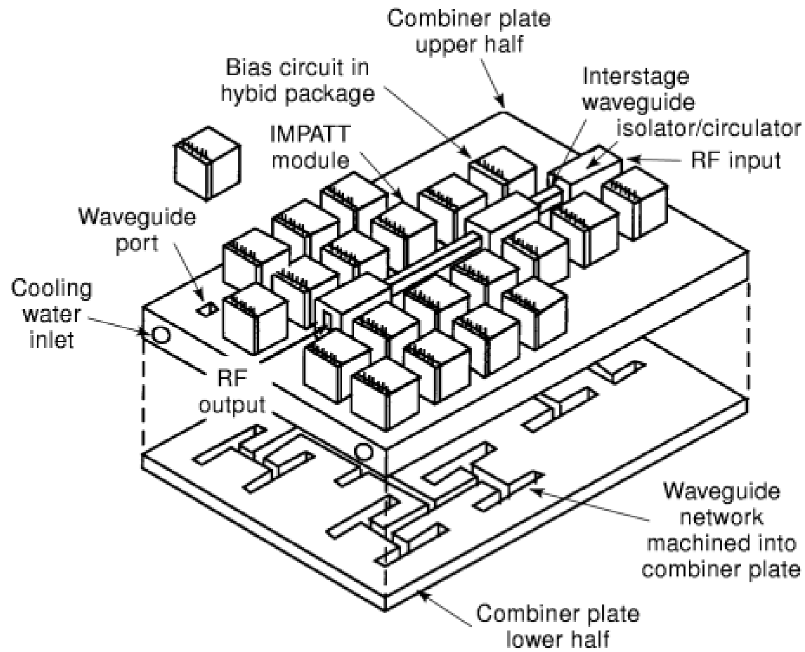

Fig. 24. Concept drawing of an IMPATT transmitter using a combiner plate.

shorts, and the data are analyzed as if the combiner were a one-diode circuit. The offset short method is also preferable for circuits that operate at high millimeter-wave frequencies where coaxial network analyzer measurements are unfeasible or insufficiently accurate. The disadvantage of the offset method is that it works only when there is good coupling between the coaxial line and the RF output port. Therefore, this method cannot be used to characterize most circuits at frequencies where parametric oscillations occur.

Transmitters Using a Combiner Plate

This article has explained the basic principles of IMPATT circuit design. As a conclusion, a new power combining technique that uses the IMPATT circuits discussed earlier will be described.

Many millimeter-wave systems require high-power transmitters, especially missile radar seekers, so an efficient means for combining hundreds of IMPATT diodes is needed. In the 1970s and 1980s, cylindrical Kurokawa power combiners were developed with as many as 60 IMPATT diodes (12). However, these circuits had limited bandwidth, and the failure of one diode would seriously detune the other devices, resulting in poor graceful degradation.

A new transmitter architecture was invented at Raytheon in the late 1980s which has been applied with great success to many frequency ranges and power levels (13,14,15). The basic concept of a combiner plate (16) is illustrated in Fig. 24. The IMPATT diodes are mounted in rectangular Kurokawa “modules,” typically with four devices per module. Many of these modules are then power combined passively by the combiner plate.

The combiner plate consists of two plates with square grooves machined on the mating surfaces. The pattern of grooves is the same on both plates, so a network of rectangular waveguides is formed when the plates are mated. The joint is in the middle of the broad wall of the waveguides where no RF current will cross the joint. This results in very low attenuation within the combiner.

The transmitter is designed to use a binary number of IMPATT modules in every stage. Magic tee junctions are built into the waveguide network to split the RF input power to the modules in a stage and then recombine their outputs. Exceptionally good port matches (typically −30 dB return loss) can be achieved with magic tees, so the modules can be integrated with the combiner plate without retuning. This is important for achieving the manufacturing cost goals for these transmitters. The magic tees also permit high isolation (typically −30 dB) between module ports, so this architecture also has very good graceful degradation.

The combiner plate is used as the main mechanical structure of the IMPATT transmitter with most components attached to it. The combiner plate is also the thermal heat sink. Water cooling passages can be machined into the plate or air cooling fins attached to the bottom. In missile applications, the thermal mass of the transmitter is adequate to maintain a safe temperature for the few seconds of operation that are required.

The combiner plate architecture has been used to develop very powerful solid state transmitters for military applications. The performance characteristics cannot be reported here, but they generate more power in the same volume as a traveling wave tube amplifier. Moreover, solid state transmitters have lower phase noise, better reliability as a result of graceful degradation, and they do not need high-voltage electronics.

Conclusions and Future Directions

IMPATT transmitters are emerging as the best technology for high-power millimeter-wave applications, displacing even traveling wave tube amplifiers. This article has given an overview of the characteristics of IMPATT diodes, the operating principles of IMPATT circuits, descriptions of the most common types of circuits, and how they can be power combined to make a high-power, multistage transmitter.

As this technology enters large-scale production, there will be pressure to reduce the cost to fabricate and assemble these transmitters. Innovations which might result are hybrid microstrip IMPATT circuits or fully integrated *MMIC* IMPATTs with their matching circuits. These new IMPATT circuits may appear to be radically different from today's technology, but they will use the same operating principles that were described in this article, only they will be applied to another transmission line medium.

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