Sensor and Simulation Notes

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Antennas for the Switched-Oscillator Source

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Abstract

This paper describes two antenna concepts for radiating a resonant waveform from a switched-oscillator source. The first has a resonant element radiating into a paraboloidal reflector. The second uses a TEM-fed paraboloidal reflector with a matched blocking circuit to isolate the slowly charged high voltage on the switched oscillator.

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1. Introduction

A recent paper [5] has considered the design of a switched oscillator as a way of generating a resonant waveform to be radiated by some antenna. Thereby one can increase the interaction with some eletronic system [3, 4]. This type of source can be thought of as a medium bandwidth source, with bandwidth somewhere between narrowband sources (microwave tubes) and hyperband sources such as impulse radiating antennas (IRAs) [6] which have typically two decades of *band ratio*.

This paper introduces some concepts for antennas appropriate to be used with switched oscillators to effectively radiate the oscillating waveform. In subsequent sections we discuss two types of such antennas. One involves a resonant radiating element feeding a reflector. The second utilizes a TEM feed to a reflector similar to that in a reflector IRA.

2. The Electric-Dipole-Fed Reflector

Figure 2.1 shows a special kind of dipole-fed reflector. The quarter-wave (in the dielectric medium) oscillator is connected to an antenna of approximately a half-wave long (in air or SF_6). The half wavelength makes the antenna element resonant at the same frequency as the oscillator. Figure 2.2 shows the approximate voltage and current distribution. In a simple (but approximate) transmission-line model the characteristic impedance Z_c of the oscillator section is much less than Z_a of the antenna section.

The oscillating antenna element might be of circular cylindrical shape from the oscillator. To better avoid electrical breakdown, this might look like an equivalent asymptotic conical dipole (ACD) [1, 2]. In addition near the connection to the source there might be some special insulating dielectric medium (e.g., oil, polyethylene) to handle the high electric fields there.

For greater fields at a distance one needs some significant antenna gain. This is efficiently accomplished by a paraboloidal reflector with focus at the dipole-like element. Noting the polarization of electric-dipole fields the reflector should not extend to the left of the xy plane (or $z = 0$ plane), at least above the dipole (near the $+y$ axis). So one might choose

$$\frac{\text{focal length}}{\text{diameter}} \equiv \frac{F}{D} = 0.25$$
$$\frac{\text{focal length}}{\text{radius}} \equiv \frac{F}{a} = 0.5 \qquad (2.1)$$

Larger values intercept less of the power radiated by the dipole, so the above may be a reasonable choice. One could extend the reflector to $+z$ values near the ground plane, and truncate top portions of the reflector (near the $+y$ axis) which do not intercept much power.

Consider now the size of the reflector as parameterized by the focal length F. Near the ground plane we can imagine an image dipole near z = -2F with opposite current (due to negative reflection at the conducting reflector). In the forward direction (+z) of the reflector beam the forward radiation from the dipole will add to the reflector beam by judicious choice for F. To make the two waves add in phase we need

$$F = \frac{2n+1}{4}\lambda$$
$$\lambda = \frac{c}{f} = \text{wavelength}$$
$$f = \text{frequency} \qquad (2.2)$$
$$c = [\mu_0\varepsilon_0]^{-1/2} = \text{speed of light}$$
$$n = \text{integer} \geq 0$$

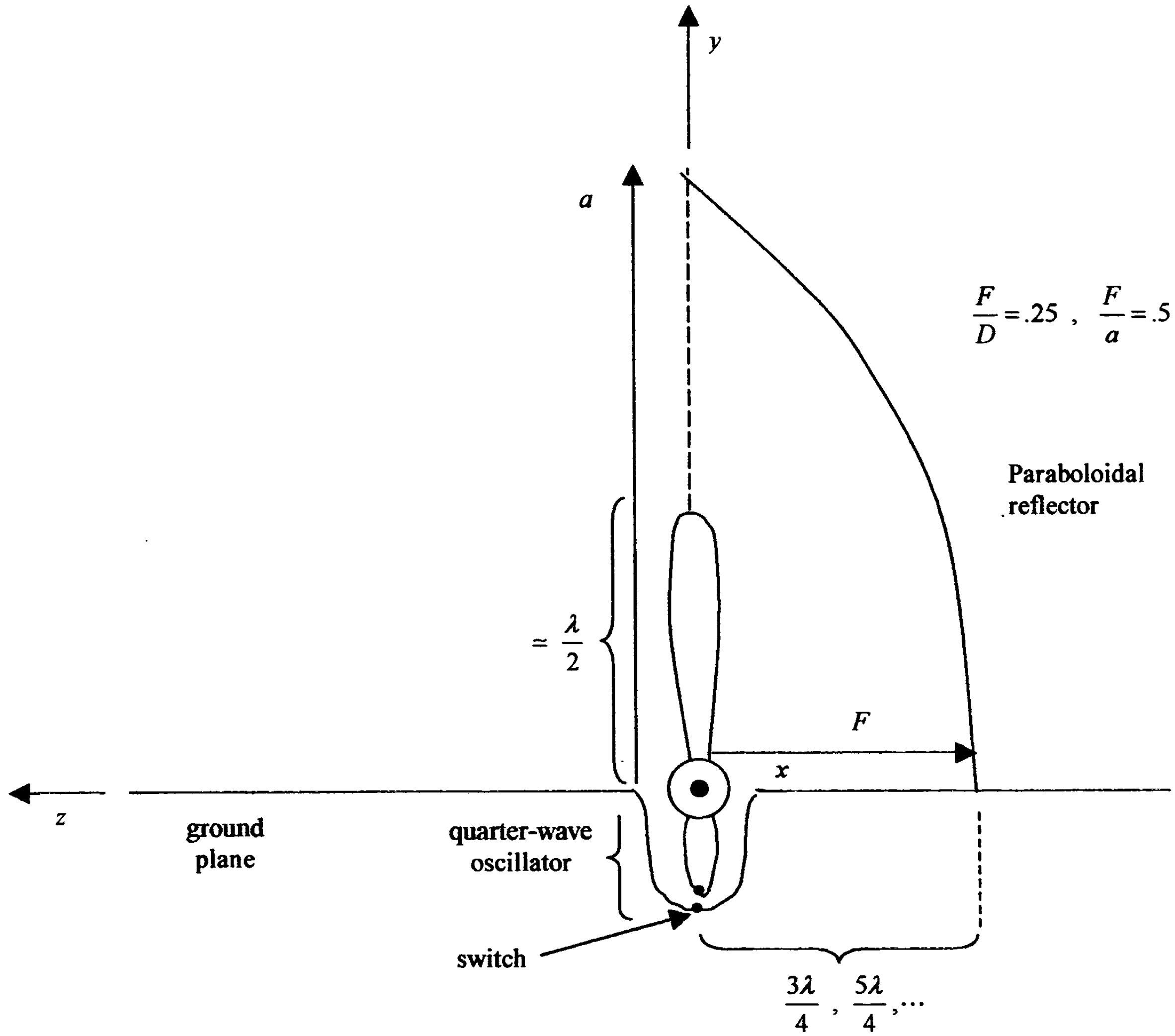

Fig. 2.1 Electric-Dipole-Fed Reflector

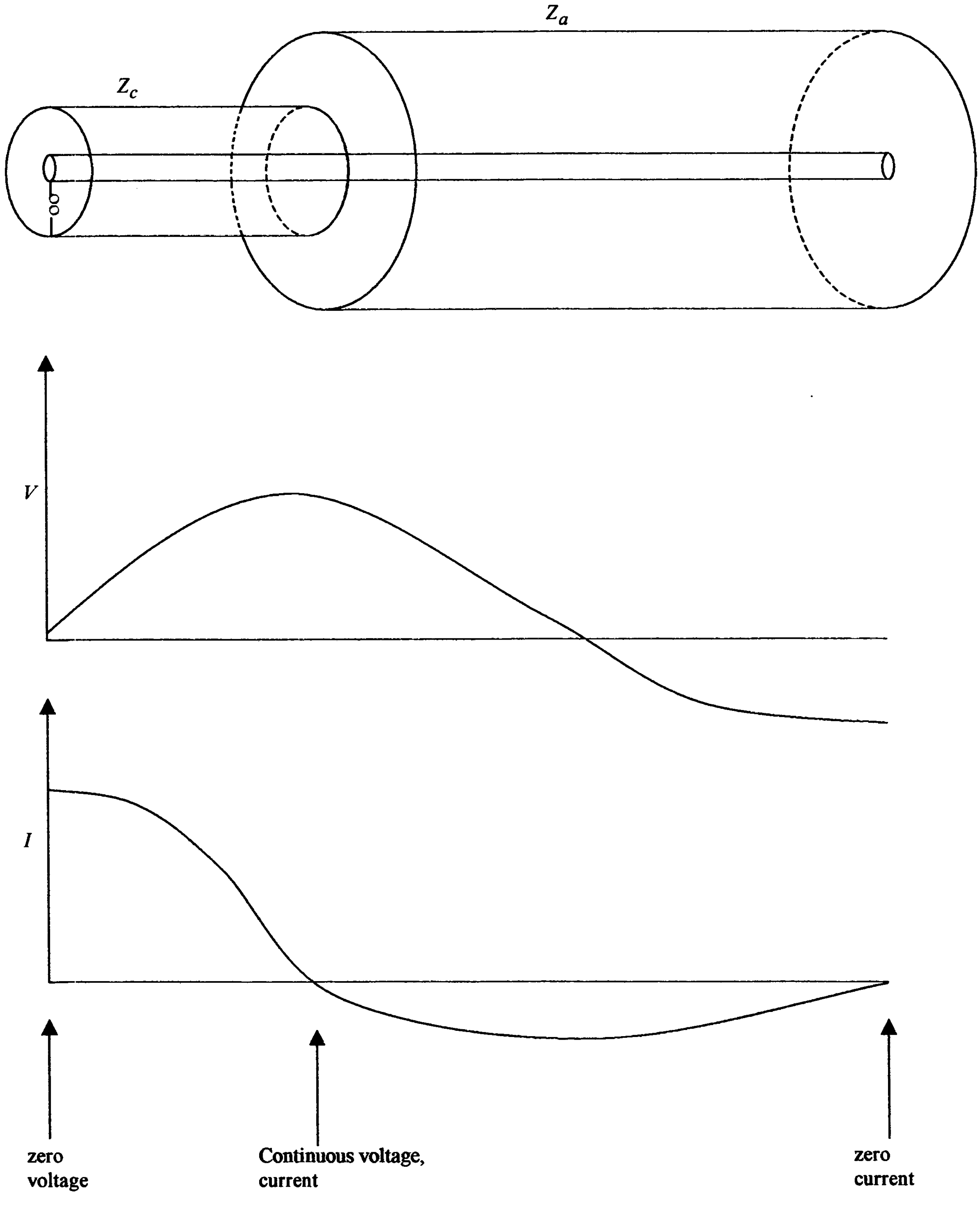

Fig 2.2 Approximate-Transmission-Line Model

Next, what is the optimal choice for n? Since $n = 0$ implies $a \simeq \lambda/2$ thereby touching the top of the dipole element (and giving only low gain). So let as constrain

$$\begin{aligned} n &\geq 1 \\ F &= \frac{3\lambda}{4}, \frac{5\lambda}{4}, \ldots \end{aligned} \tag{2.3}$$

This gives only a rough description. The details of the switched oscillator and dipole element need to be considered for accurately matched resonant frequencies. This in turn will affect the reflector design. A limitation of this type of antenna is the low radiation efficiency of a small antenna element. Note also that the antenna input impedance is not a simple constant resistive impedance such as used in [5].

Then there are the high voltage and switching considerations. In particular, the reflector can be part of a barrier to contain SF6. Near the base of the antenna element the insulating gas may not be adequate, and an oil region here may be appropriate.

3. TEM-Fed Reflector with Blocking Capacitor

Consider the antenna in Fig. 3.1. It has a TEM feed of characteristic impedance Z_f, typically using two feed arms giving an impedance in the 100 Ω ballpark. With the apex of this conical transmission line at the apex of a paraboloidal reflector, this has some similarity to the reflector IRA. Since we are considering concentrating the operation around some radian frequency $\omega_0 = 2\pi f_0$ we can revisit the choice of the terminating impedance $\tilde{Z}_t(s)$ for optimum results. Note that $\tilde{Z}_t(s)$ is the parallel combination of the two terminating impedances connecting the two feed arms to the reflector. Here we use the two-sided Laplace transform with

$$\tilde{f}(s) = \int_{-\infty}^{\infty} f(t) e^{-st} dt$$

$$f(s) = \frac{1}{2\pi j} \int_{Br} \tilde{f}(t) e^{st} dt$$

$$s \equiv \Omega + j\omega \equiv \text{Laplace-transform variable or complex frequency} \tag{3.2}$$

$$\gamma_d = \frac{s}{v} \;,\; v = \text{propagation speed in dielectric medium}$$

Near the output of the switched oscillator there is a blocking capacitance C (parallel combination of two in the arms). This has the function of isolating the high voltage source from the antenna. Of course the source must now charge $C_0 + C$ where C_0 is the oscillator capacitance. So we need to choose C carefully. The charging current through C returns through the feed arms and reflector to the ground plane so $\tilde{Z}_t$ must be finite at low frequencies (perhaps even a short circuit).

A disadvantage of a blocking capacitor is the impedance it presents to the oscillator, lowering the voltage delivered to the feed arms. One way to compensate for this is to include a series inductance L (net parallel combination) such that, at the frequency of interest, the series combination has zero impedance $\tilde{Z}_b(s)$, i.e.,

$$\begin{aligned} \tilde{Z}_b(s_b) &= 0 = \frac{1}{s_b C} + s_b L = \frac{1}{j\omega_b C} + j\omega_b L_b \\ \omega_b &= [LC]^{-\frac{1}{2}} \;,\; L = \left[\omega_b^2 C\right]^{-1} \end{aligned} \tag{3.2}$$

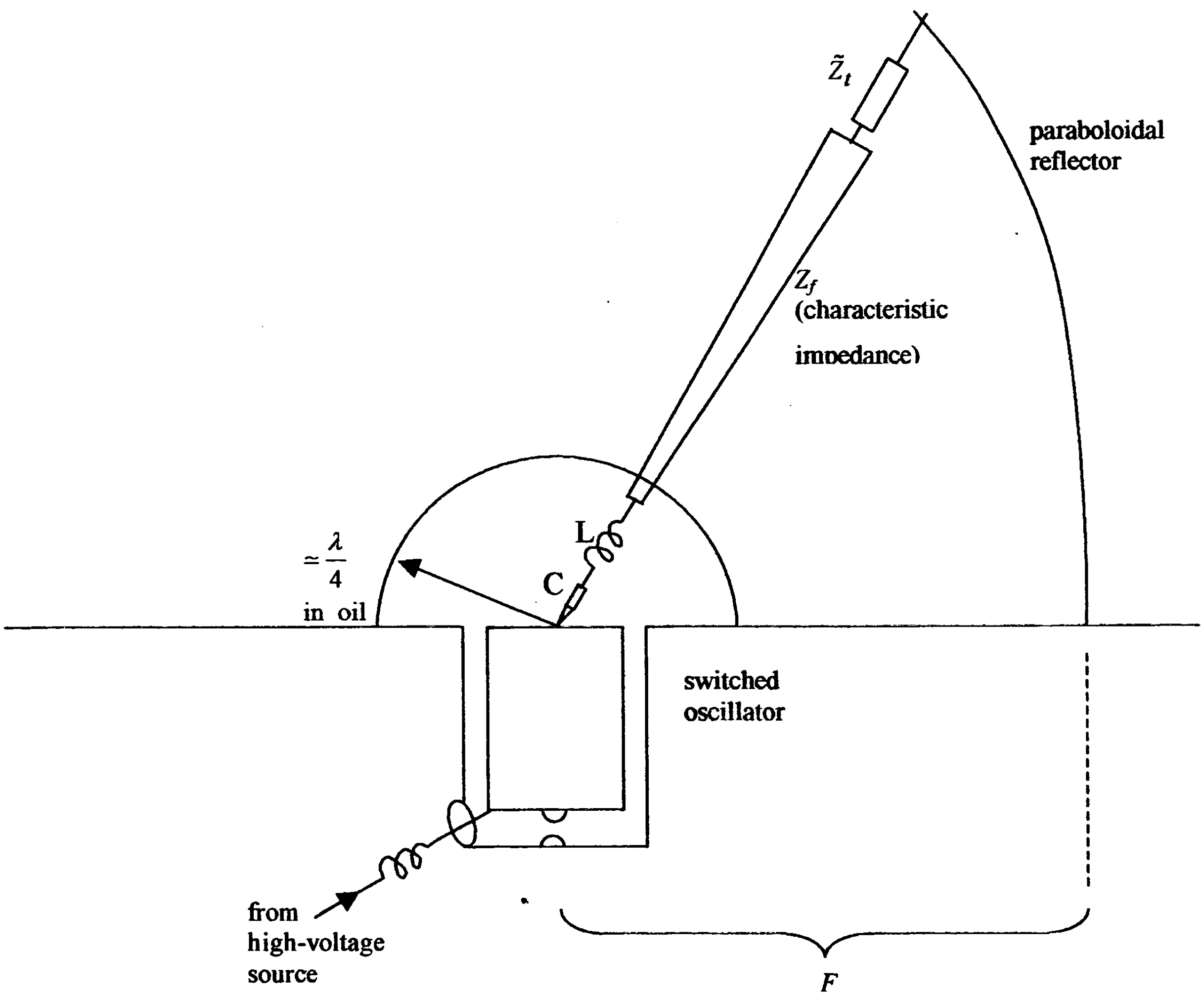

Fig. 3.1 TEM-Fed Reflector

Of course, a pulse contains a band of frequencies and the match is at what might be regarded as the center frequency. Note that series resistance can also be included in the above, if desired.

The impedance now loading the switched oscillator is $Z_c + \tilde{Z}_b$. The equivalent circuit is as illustrated in Fig. 3.2. Note for the moment that the hemispherical dielectric region containing $\tilde{Z}_b$ (for high-voltage reasons) is ignored. This might be chosen as a quarter-wavelength in the dielectric at the frequency of interest for improved matching to the antenna. Modelling the switch closure by a voltage step $V_0 u(t)$ we need to optimize $V_a(t)$ in some sense as the waveform delivered to the antenna feed. (This can be added to the initial condition of $-V_0$ on the oscillator, if desired.)

Defining a normalized impedance

$$\tilde{\zeta}(s) \equiv \frac{Z_f + \tilde{Z}_b(s)}{Z_c} \tag{3.3}$$

we have a reflection coefficient at the end of the oscillator ($z = \ell$) as

$$\tilde{\xi}(s) = \frac{\tilde{\zeta}(s)-1}{\tilde{\zeta}(s)+1} \tag{3.4}$$

The voltage waves on the oscillator combine as

$$\tilde{V}(z,s) = \tilde{V}_1 e^{-\gamma_d z} + \tilde{V}_2 e^{\gamma_d z} \tag{3.5}$$

with boundary conditions

$$\begin{aligned} \tilde{V}(0,s) &= \frac{V_0}{s} = \tilde{V}_1 + \tilde{V}_2 \\ \tilde{V}(\ell,s) &= \tilde{V}_1 e^{-st_t} + \tilde{V}_2 e^{st_t} \end{aligned} \tag{3.6}$$

At $z = \ell$ the left-going wave is $\tilde{\xi}$ times the right-going wave giving

$$\tilde{V}_2 e^{st_t} = \tilde{\xi}(s)\tilde{V}_1 e^{-st_t} \tag{3.7}$$

Combining we have

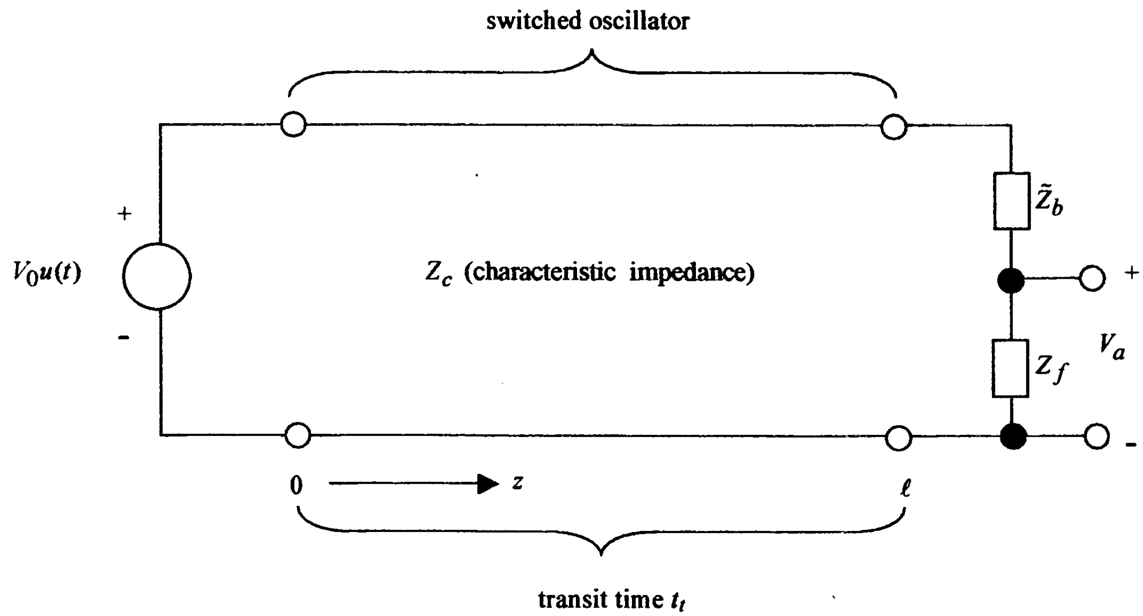

Fig. 3.2 Equivalent Circuit of Source and Antenna

$$\frac{\tilde{V}(\ell,s)}{\tilde{V}(0,s)} = \frac{\left[1 + \tilde{\xi}(s)\right]e^{-st_t}}{1 + \tilde{\xi}(s)e^{-st_t}} \tag{3.8}$$

describing the switched oscillator.

At the input to the antenna feed we have a voltage divider as

$$\frac{\tilde{V}_a(s)}{\tilde{V}(\ell,s)} = \frac{Z_f}{Z_f + \tilde{Z}_b(s)} = \frac{Z_f}{Z_c}\tilde{\zeta}^{-1}(s) \tag{3.9}$$

Combining with the previous we have

$$\frac{\tilde{V}_a(s)e^{st_t}}{V_0} = \frac{1}{s\tilde{\zeta}(s)}\,\frac{Z_f}{Z_c}\,\frac{1+\tilde{\xi}(s)}{1+\tilde{\xi}(s)e^{-2st_t}} \tag{3.10}$$

The inverse transform of this gives the antenna voltage in retarded time.

The special case of $\tilde{Z}_b = 0$ is given by

$$\frac{\tilde{V}_a(s)e^{st_t}}{V_0} = \frac{1}{s}\,\frac{1+\tilde{\xi}(s)}{1+\tilde{\xi}(s)e^{-2st_t}}$$
$$\tilde{\xi}(s) = \frac{\dfrac{Z_f}{Z_c} - 1}{\dfrac{Z_f}{Z_c} + 1} \tag{B.11}$$

The time domain form of $V_a(t-t_t)$ is discussed in some detail in [5]. For $Z_f / Z_c >> 1$ the peak value is almost $2V_0$. The complete waveform is a sequence of steps with decreasing amplitudes — sort of a "square" damped sinusoid.

Reducing the form in (3.10) we have

$$\frac{\tilde{V}_a(s)e^{st_t}}{V_0} = \frac{2}{s}\frac{Z_f}{Z_c}\left[\left[\tilde{\zeta}(s) + 1\right] + \left[\tilde{\zeta}(s) - 1\right]e^{-2st_t}\right]^{-1}$$
$$= \frac{2}{s}\left[\left[1 + \frac{\tilde{Z}_b(s) + Z_c}{Z_f}\right] + \left[1 + \frac{\tilde{Z}_b(s) - Z_c}{Z_f}\right]e^{-2st_t}\right]^{-1} \tag{3.12}$$

Looking at the resonant behavior we need the zeros of the denominator. For the special case of $\tilde{Z}_b = 0$ we have the oscillator complex resonance as

$$-2s_0t_t = \ell n\left(-\frac{1 + \frac{Z_c}{Z_f}}{1 - \frac{Z_c}{Z_f}}\right)$$
$$s_0t_t = [\Omega_0 + j\omega_0]t_t = -\frac{1}{2}\ell n\left(-\frac{1 + \frac{Z_c}{Z_f}}{1 - \frac{Z_c}{Z_f}}\right) \pm j\frac{\pi}{2} \tag{3.13}$$
$$\simeq -\frac{Z_c}{Z_f} \pm j\frac{\pi}{2} \text{ as } \frac{Z_c}{Z_f} \to 0$$
$$f_0 = \frac{|\omega_0|}{2\pi} = \frac{1}{4t_t}$$

Note that ω_0 takes both signs to account for the conjugate pair. With $4t_t$ as a period, the number of cycles to e^{-1} is just $Z_f/(4Z_c)$ as in [5].

If we modify $\tilde{Z}_b$ in (3.2) to give some series loss of an amount to make $\tilde{Z}_b(s_0) = 0$, this simplifies the analysis. Then the denominator in (3.12) is still exactly zero at the oscillator complex frequency. Then we need the derivative of this function at s_0 to obtain the residue. We have

$$\frac{d}{ds}\left[1+\frac{\tilde{Z}_b(s) + Z_c}{Z_f}\right]\Bigg|_{s=s_0} = \frac{1}{Z_f}\frac{d}{ds}\tilde{Z}_b(s)\Bigg|_{s=s_0}$$
$$= \frac{1}{Z_f}\left[-\frac{1}{s_0^2C} + L\right] = \frac{2L}{Z_f} = \frac{2}{|\omega_0|Z_f}\left[\frac{L}{C}\right]^{\frac{1}{2}} \equiv \frac{\chi}{|\omega_0|}$$
$$\chi \equiv \frac{2}{Z_f}\left[\frac{L}{C}\right]^{\frac{1}{2}}$$

$$\frac{d}{ds}\left[\left[1+\frac{\tilde{Z}_b(s)-Z_c}{Z_f}\right]e^{-2st_t}\right]\Bigg|_{s=s_0}$$

$$=-2t_t e^{-2s_0t_t}\left[1-\frac{Z_c}{Z_f}\right]+e^{-2s_0t_t}\,\frac{\chi}{|\omega_0|} \tag{3.14}$$

Then we approximate as

$$e^{-2s_0t_t} = -\frac{1+\frac{Z_c}{Z_f}}{1-\frac{Z_c}{Z_f}} \simeq -\left[1+2\frac{Z_c}{Z_f}\right]$$

$$\frac{\tilde{V}_a(s)\,e^{st_t}}{V_0} \simeq 2\left[\frac{\chi}{|\omega_0|}+\left[2t_t\left[1-\frac{Z_c}{Z_f}\right]-\frac{\chi}{|\omega_0|}\right]\frac{1+\frac{Z_c}{Z_t}}{1-\frac{Z_c}{Z_t}}\right]^{-1}$$

$$\left[\frac{1}{s_0}[s-s_0]^{-1}+\frac{1}{s_0^*}\left[s-s_0^*\right]^{-1}\right] \tag{3.15}$$

$$\simeq 2\left[\chi+\left[\pi\left[1-\frac{Z_c}{Z_f}\right]-\chi\right]\left[1+2\frac{Z_c}{Z_f}\right]\right]^{-1}$$

$$\left[\frac{|\omega_0|}{s_0}[s-s_0]^{-1}+\frac{|\omega_0|}{s_0^*}\left[s-s_0^*\right]^{-1}\right]$$

including only the dominant pole pair.

Converting this to time domain we have

$$\frac{V_a(t)}{V_0} \simeq 2\left[\chi+\left[\pi\left[1-\frac{Z_c}{Z_f}\right]-\chi\right]\left[1+2\frac{Z_c}{Z_f}\right]\right]^{-1}$$

$$\left[\frac{|\omega_0|e^{s_0[t-t_t]}}{s_0}+\frac{|\omega_0|e^{s_0^*[t-t_t]}}{s_0^*}\right]u(t-t_t) \tag{3.16}$$

$$=4\left[\pi+\frac{Z_c}{Z_f}[\pi-2\chi]\right]^{-1}e^{\Omega_0[t-t_0]}\sin(|\omega_0|[t-t_t])\,u(t-t_t)$$

For further insight let $Z_c/Z_f \to 0$ giving

$$\begin{aligned}\Omega_0 &\simeq 0 \\ \frac{V_a(t)}{V_0} &\simeq \frac{4}{\pi}\sin\left(|\omega_0|\left[t-t_t\right]\right)u\left(t-t_t\right)\end{aligned} \tag{3.17}$$

which has amplitude $4/\pi$ which is the factor relating a square wave oscillating between +1 and −1 (or equivalently +2 and 0) to the fundamental sinusoidal component.

If $Z_c / Z_f > 0$ then (3.16) shows some effect of the blocking network through the parameter χ. From (3.10) this can be considered as some kind of normalized characteristic impedance of the blocking network. While one may desire small C to minimize loading the high-voltage source, a given ω_0 will imply a correspondingly large L and χ, thereby eventually making the effect of the blocking circuit or the amplitude of the basic resonance significant.

While the present analysis is approximate, one can return to (3.12) for a more accurate version of the Laplace transform, and via a numerical inverse transform obtain the time-domain waveform. Also one can vary the form of $\tilde{Z}_b$, e.g., by removing the inductance. Alternately one can do a more complete pole expansion, including not only the domainant oscillator resonance (perhaps shifted).

4. Concluding Remarks

These then are two concepts for an antenna matched to the switched oscillator. Each has its special characteristics. Reviewing the various design considerations one can see that there are various options and details to consider. Not only are there electromagnetic considerations, but high-voltage ones as well (e.g., switching and breakdown of various materials).

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